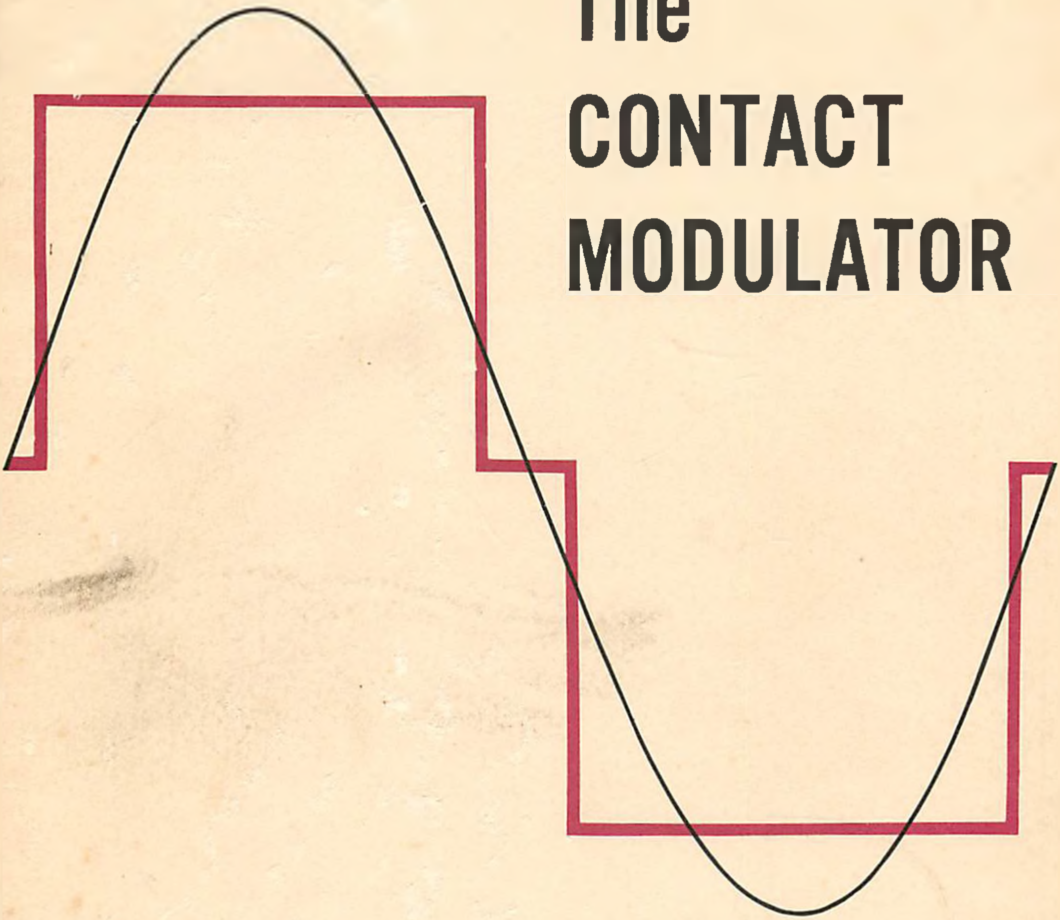
The CONTACT MODULATOR



PART 1: Why Use Choppers?

THE FIRST OF A SERIES

AIRPAX

THE

CONTACT MODULATOR

PART I

WHY USE CHOPPERS?

WRITTEN BY

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and

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Cambridge, Maryland • Fort Lauderdale, Florida
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Price One Dollar

FOREWORD

The Chopper, sometimes referred to as a contact modulator, occupies a unique position in modern circuitry. It is the connecting link between low or zero frequency, low level transducer and the amplifying system designed to increase transducer output to a usable value.

This first part is concerned with definitions, techniques of measurement, general usage and standards of test and terminology.

Airpax was instrumental in coining and bringing into accepted usage many of the terms now applied internationally to chopper technology.

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SECTION I

WHY USE CHOPPERS?

The name chopper is derived from the contact devices used to tone-modulate CW when CW first replaced spark transmission. The contact modulator or chopper came into widespread use about 1948, with the appearance of a wide variety of servo-mechanisms and DC amplifiers.

There ought to be something wrong with using electromechanical devices to perform electronic functions. Or at least, this seems obvious to many of us for whom moving devices are anathema. Also, there are several other ways of performing the chopper function; and, it is fair to ask, why use choppers at all?

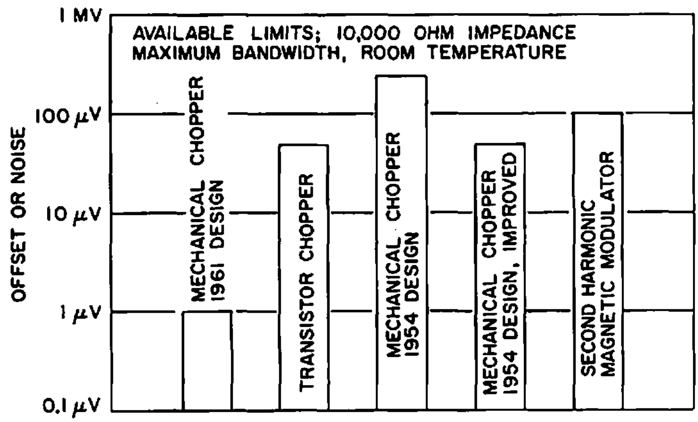


FIGURE 1. *Null stability of various modulators.*

Magnetic modulators, for example, provide probably the most reliable substitute, the second harmonic type having considerably better zero stability than the fundamental type. Figure 1, compares various modulator types. The bar chart compares three kinds of modulators. It makes the assumption that the circuit impedance is 10,000 ohms, that the system bandwidth is as wide as possible, that we consider only room temperature, that we ignore size and ignore expense. Finally, it tries to compare choppers (a component) with a magnetic modulator (a little system).

With a second harmonic type magnetic modulator, under controlled laboratory conditions, we can reach a null stability as low as 10 microvolts—but—the response time is less than 3 CPS, the impedance rather low, the ambient temperature range is severely restricted, and the modulator is larger, heavier and more expensive. At impedances of 100 ohms, 20 to 30 uv is a practical lower limit.

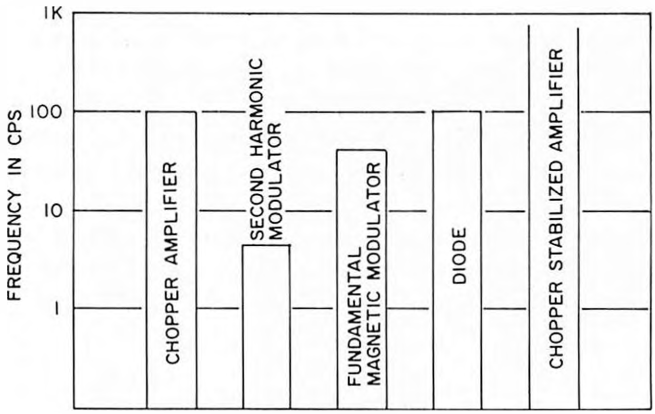


FIGURE 2. *Frequency response limits of modulator systems. A 400 cycle carrier is assumed.*

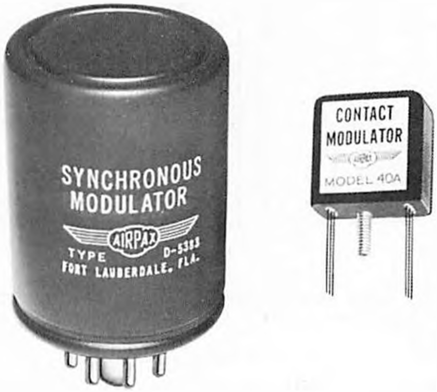


FIGURE 3. *Airpax magnetic modulator compared to Airpax chopper.*

The chart in Figure 2, assuming a carrier frequency of 400 CPS, shows the frequency response limits of various modulation methods.

We can make trades to improve a desired parameter of course, but

the response is limited by the time constants of the input winding and input resistance. Better frequency response is possible, at the expense of impedance or gain. Figure 3, compares a magnetic modulator to a chopper.

The best in transistor choppers is depicted by Airpax Type 7000, which exhibits a remarkable degree of isolation between drive and modulation elements. In consequence, the noise level is down to about 35 uv at relatively high circuit impedances of approximately 10,000 ohms.

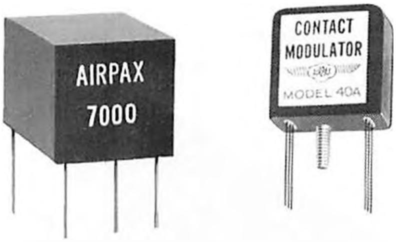


FIGURE 4. *Transistor chopper and a contact modulator.*

FIGURE 5. *Airpax Type AMS 34 DC Amplifier.*

Magnetic and semiconductor components are considered solid state devices. Magnetic amplifiers in particular, are stable and capable of very long life. Transistor amplifiers are light, compact and require very low power. Airpax produces magnetic amplifiers and transistor amplifiers, as well as choppers. Where these devices can be applied and continuous unattended service for years is a requirement, the magnetic amplifier would be the logical choice.

There are other modulation devices, photo-electric, tube oscillator, etc. Actually, the situation which arises is that the stability which can be achieved with good DC amplifiers (without choppers) is about as good as can be reached with other modulation methods.

It is now practical to design an unstabilized DC transistor amplifier, solid state, which will hold to 1 or 2 millivolts at the output with a closed loop gain of 10, at room temperature, and 3 to 4 mv over a wide temperature range. The Airpax AMS 34 is such an amplifier, Figure 5, and is a good example of what is practical today without chopper stabilizing. As an operational amplifier with a gain of 10, the response at 3 db down is about 5000 cycles. Better stability will demand choppers. At a bandwidth of 0 to 30 KC, a null stability of about 2 millivolts is possible over about 50°C and a gain of 10, but more expensive components are required. Such an amplifier is Airpax Type AMS 26, illustrated in Figure 6.

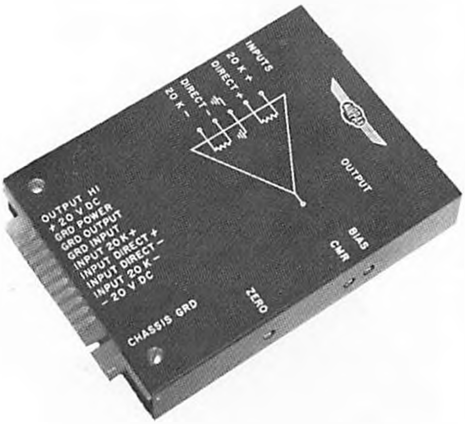


FIGURE 6. *Airpax Type AMS 26 DC Amplifier.*

The use[1] of a stabilizing chopper or of a carrier system becomes necessary when: the gain must be greater than 10; still wider frequency ranges become necessary; very low signals in the microvolt range must be amplified; high impedance is needed; or when expense is important. It is possible to improve performance somewhat beyond the DC amplifier performance given, using several different modulator systems (thus a transistor chopper will improve a transistor amplifier), but when the limits of the state of the art must be reached only a mechanical chopper will do.

Why? Well—it's fairly simple.[2] Contact modulators are unique in one special respect, their ability to switch between a very low resistance of much less than 0.1 ohm to over 10,000 megohms, and in very short time intervals.

[1]On the Design of Chopper Stabilized Amplifiers, The Contact Modulator Part III.
[2]A Study of the Transfer Function of Contact Modulated Amplifiers. Krantz, Salati and Berkowitz, AIEE Transactions, Paper No. 57-204.

When should a chopper NOT be used? Obviously, the decision to use a chopper, or not to use one, can only be made with a full knowledge of the circuit requirements, however, some generalities can be discussed.

If life requirements for a piece of equipment call for over 10,000 hours of unattended use, a chopper may not be suitable, or at least, one may have trouble convincing people a chopper can be used. Choppers are supplied in plug-in types and can be replaced as easily as a vacuum tube or similar plug-in component. Choppers have been observed to have a life that evidently exceeds 25,000 hours, or roughly three years of continuous service. The difficulty of extensive life testing is obvious. At present, Airpax rates Type 300, on which the most data is available, at 10,000 hours. If an exceedingly long life is a must, a magnetic amplifier, such as an Airpax Ferrac or Preac, should be used. These devices have an expected life of 100,000 hours.

FIGURE 7. *Magnetic amplifier for temperature system.*

Figure 7 shows such an amplifier designed for a maximum of life, operating from an undersea temperature probe to obtain synoptic data of ocean temperature.

If high input impedance and good null stability are required of a DC amplifier, a chopper may be the only practical solution. At impedances of around 1,000 ohms, and with not too much of a drift problem, diode or transistor modulators can be used. In general, if good zero stability is needed with nulls less than about 35 microvolts and impedances over a few hundred ohms, a chopper is almost imperative. (However, it is also quite practical with modern techniques to get pretty fair performance without a modulator.)

The response time needed enters the problem. "Chopper Amplifier" refers to an amplifier using a chopper to modulate the input DC; probably also to demodulate. The response is limited to about 1/4 the driving frequency. "Chopper Stabilized Amplifier" means stabilizing the gain of a DC amplifier with a chopper amplifier—in this case the possible response can be up to a very high frequency. The term "DC Amplifier" means of course, any amplifier having response to zero frequency.

The chart shown in Figure 1, described the minimum practical offset obtainable from various modulator types. No such chart can present a full picture, obviously there are limits such as frequency response, circuit impedance and temperature range. One of these limits was found in Figure 2. And, since these are minimums, the limits also depend on other engineering interpretations. The closest approach to chopper stability, the second harmonic magnetic modulator, is limited by slow response, about 1 to 3 CPS; by temperature changes, which give poorer stability; and by the second harmonic in the power source, which causes offset.

The first question therefore, is whether or not a chopper is needed. Other than modulation of DC, you may want to compare two DC signals, to switch radio frequency signals, to provide a time sharing method, to stabilize another amplifier, to provide a filter circuit, to demodulate or rectify signals, or to sample data. With a problem in the zero frequency region, a mechanical chopper is required if you need a combination of these factors:

a. High impedance circuits, over 1,000 ohms.
b. Response time greater than 2-10 CPS.
c. Null stability better than 50 microvolts.
d. Or null stability better than 100 microvolts under wide temperature ranges.
e. Performance over about 85°C.
f. You are limited by space and weight.
g. You are limited by your budget.
h. Limited power.
i. Your system is exposed to atomic radiation.
j. The signal to be modulated has a very wide dynamic range.
k. Extreme linearity is demanded, as for digital conversion.
l. 10,000 hours life expectancy is sufficient.

It is probably worth pointing out that a mechanical chopper is by far the cheapest way of getting good null stability; and they were considered reliable enough to fly with John Glenn on our first manned space flight. Even if choppers are avoided altogether by careful design, for moderately good stability, the cost of the required components considerably exceeds the cost of the chopper and its entire system requirements.

In both modulation and demodulation systems, there arises a need for high stability and freedom from zero shift, or that characteristic which results in an output signal when no input signal is present. The contact device is unique in that it solves the problem of a stable minimum resistance by being close to zero when closed, and presents a maximum resistance which approaches infinity when open. And, in addition, it can be used effectively at both low and high impedances.

Choppers suffer from one problem—there is some point at which the contacts begin to wear. Ten thousand hours is a reasonable life expectancy for a properly used chopper. We may expect more than that in some circuits. Figure 8 shows internal and external construction of Airpax choppers, Types 300, 370 and 40A.

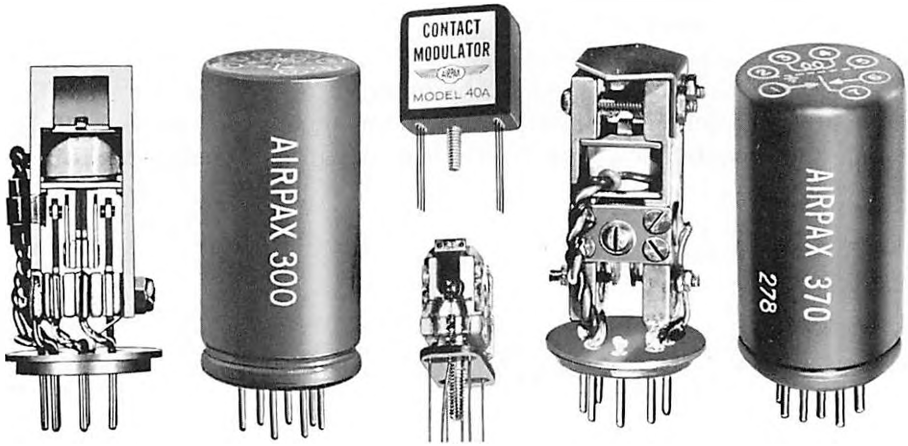


FIGURE 8. *Internal construction of Airpax standard and balanced armature choppers.*

There is an effective phase lag between contacts and driving signal which may be between 20 and 100 electrical degrees, depending on frequency and other factors. The lag is made up of two parts; lagging coil current and a lag affected by mechanical constants. The majority of Airpax choppers have a SPDT BBM (break-before-make) contact arrangement. The construction of the chopper is such that there is no neutral position of the moving arm when the chopper is not being driven, and the movable contact will stop at random on either fixed contact when the drive is removed. Contacts on the larger choppers carry a maximum voltage rating of 100 volts and a maximum current rating of 2 milliamperes, for unity power factor load, and are driven from the supply mains which are most frequently 400 CPS or 60 CPS.

In summary, a mechanical chopper is an economical way of getting high performance out of DC amplifiers. In fact, if you must have the very best performance from your circuit, you have no choice left. Only a mechanical chopper provides at one time the lowest stray noise, the ability to work over environmental changes (as temperature, shock and vibration), the best response time, the simplest circuit, the smallest size and the least weight. The dollar savings are a bonus, and you may have to put up with life expectation of only 10,000 hours.

If you understandably prefer solid state and can tolerate a null stability (referred to the input) of perhaps 100 microvolts, and you must save weight and size, a transistor chopper similar to Airpax Type 7000 can handle the application. Quite a few transistor switches are available with less performance, however as noted earlier, when you can accept null errors as high as 1 to 10 microvolts, and there are many such applications, you probably do not need a chopper at all. You can design some self balancing into your DC amplifier and the result is likely to be almost as satisfactory.

When slow response time can be tolerated, as slow as one second or an appreciable portion thereof, a magnetic modulator will provide a high order of life expectancy, if of conservative design and quality production.

SECTION II
WHICH CHOPPER?

Mechanical choppers, as normally available, are of two general types. Both use some form of polarization, usually a permanent magnet, are essentially low Q devices, and therefore will follow a driving frequency. The difference lies in the frequency of self resonance.

In Figure 8, Types 300 and 370 have a resonance at about 600 cycles, at which point there is a rapid increase in phase after which the chopper

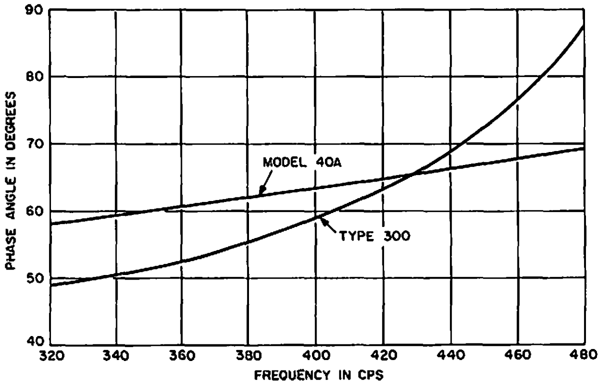


FIGURE 9.
Phase angle versus frequency of Type 300 and Model 40A.

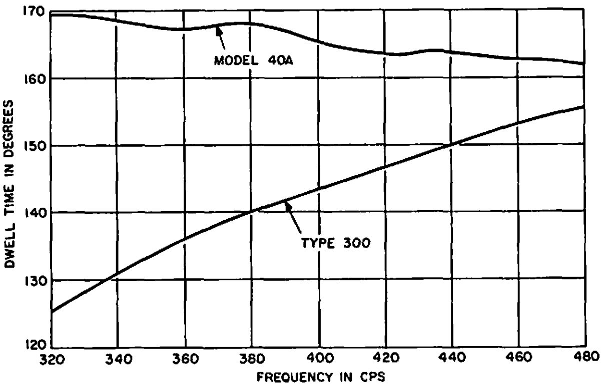


FIGURE 10.
Dwell time versus frequency of Type 300 and Model 40A.

stops operating. Type 300 pivots from one end, Type 370 is a balanced structure. These types are similar in performance. They offer high contact pressure, a high order of reliability and the benefits of many years of production. They are quite a bit larger, of course, but are also quite cheap and easily obtained from several companies. Other than size, they suffer from the fact that their phase angle is not constant with frequency, Figure 9. A side effect is a change in dwell time with frequency, Figure 10.

Model 40A differs, and can be considered as a rotating device with contacts at right angles to the rotation. It has a self resonant point of several thousand cycles, and as shown in Figure 9, does not exhibit as much

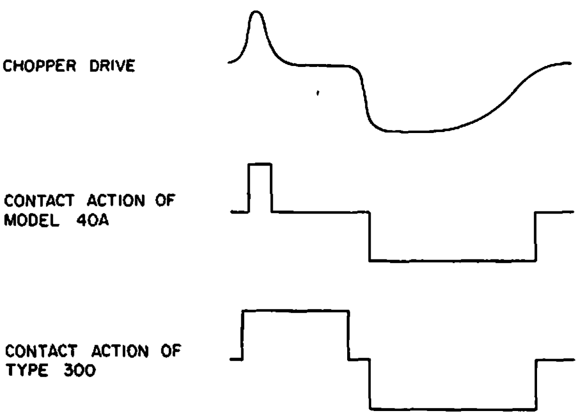


FIGURE 11. *The effects of asymmetrical drive.*

phase change with frequency. Contact pressures are lower, the relative dynamic contact travel is much less. The techniques to insure reliability, such as the use of dust free assembly rooms, are of necessity more sophisticated.

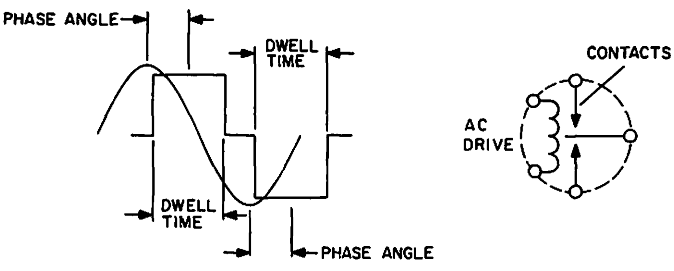


FIGURE 12. *Sine wave drive produces modified square wave when using SPDT BBM chopper.*

The mechanical motion of a chopper whose mechanical resonant period is close to its excitation frequency is approximately sinusoidal with time. As might be expected, an armature resonant at a frequency much higher than the driving signal tends to more closely follow the current wave shape of the drive.

This can be used beneficially, as when a non-symmetrical contact dwell is desired. Here if the drive signal is of shorter duration on one-half the wave, the contact dwell time will also be shorter. On the other hand, when a square wave drive is used, this may not be desirable and here the use of a chopper such as Type 300 permits ignoring the harmonic content of the drive, Figure 11.

A chopper is best defined as being driven by a sine wave and delivering a modified square wave, as in Figure 12. Since you probably want to drive from the power line, you will in all likelihood have either 6.3 or 110 volts available, at either 60 or 400 CPS.

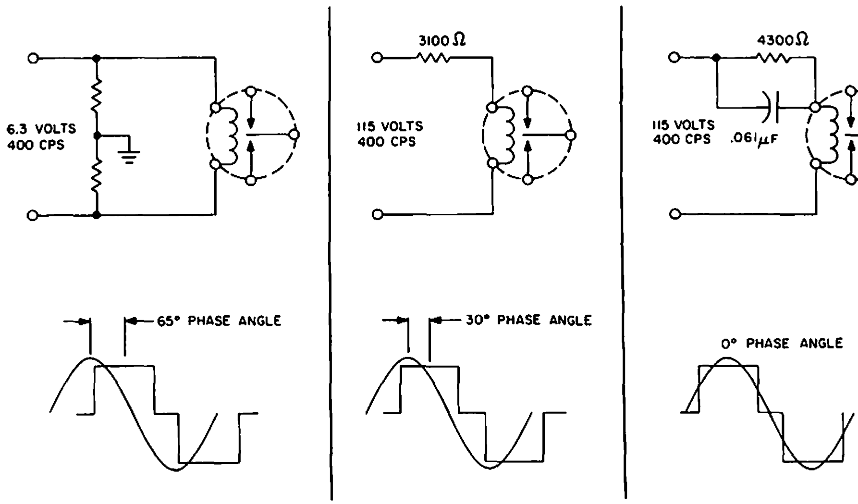


FIGURE 13. ***Type 300 chopper with 6.3 volt drive and 65° phase angle can be adapted for 115 volt drive and will produce other phase angles.***

Figure 13 illustrates a Type 300 chopper, SPDT, break-before-make, 65° phase lag, 147° dwell time, with a 6.3 volt 400 cycle drive, arranged for three different drive methods. This is one of the choppers shown in Figure 8, a popular stock model.

When the power line is not available the chopper can be driven from a pair of transistors arranged in an inverter circuit, Figure 14. This is sometimes done to provide a 400 cycle drive, even when 60 cycles is available, since a higher frequency improves the response time of a carrier system.

Type 175, illustrated in Figure 15, and Type 300, are identical in external appearance. These two types are used in many chopper amplifier applications, and hundreds of thousands of them are in service. These

types mount in a seven pin socket, JAN TS102P01, with a miniature tube shield, JAN TS102U02, securing the chopper in the socket.

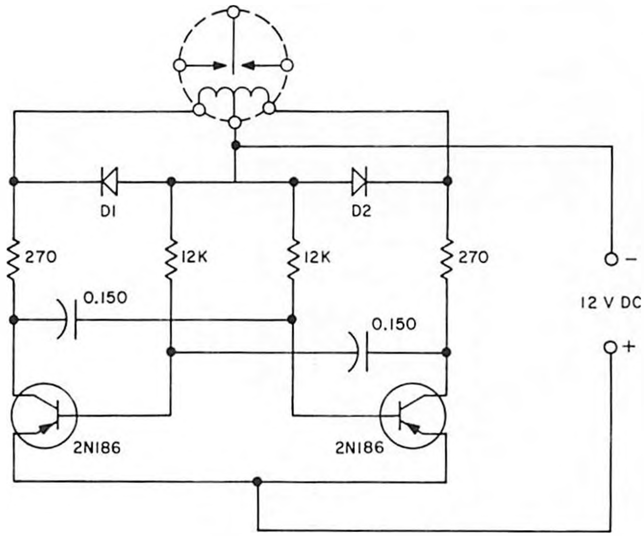


FIGURE 14. *Inverter circuit using the chopper coil.*

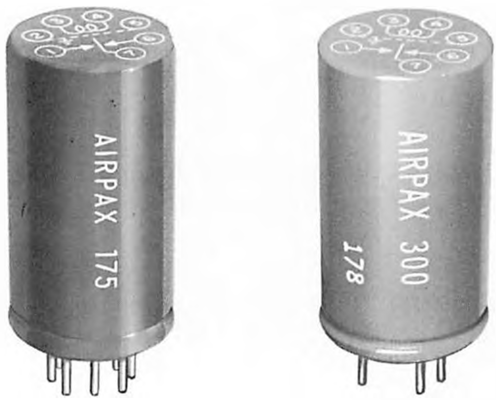


FIGURE 15. *Type 175 chopper uses a 6.3 volt 60 CPS drive, has a phase angle of 20°, and a dwell time of 164°.*

Life figures of 10,000 hours, or mounting and space considerations, may make it undesirable to plug in the chopper. Often it is helpful to solder connections directly, avoiding low level thermal junctions which may appear in sockets. Figure 16 shows a variety of mounting possibilities.

None of the standard types of Airpax choppers will be damaged if subjected to vibration between 10 and 3000 cycles, or to white noise. Where the chopper must operate disturbed by severe conditions of vibration, Type 370 may be used. See Figure 17.

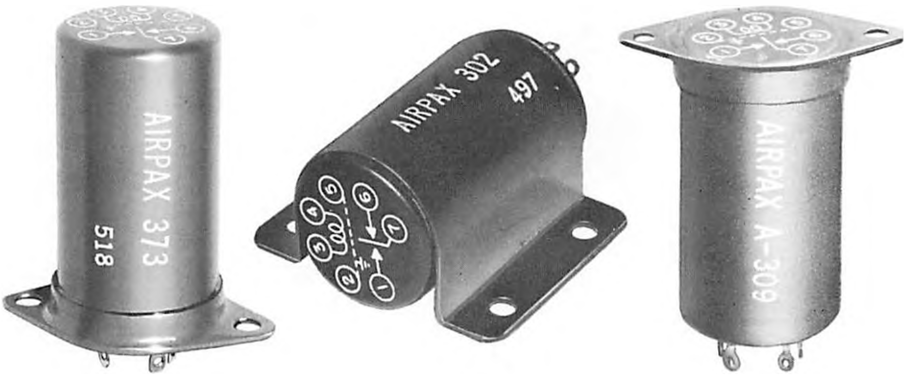


FIGURE 16. *Choppers are supplied in a variety of mountings for permanent connection. Bracket, top and bottom flange mounts are shown.*

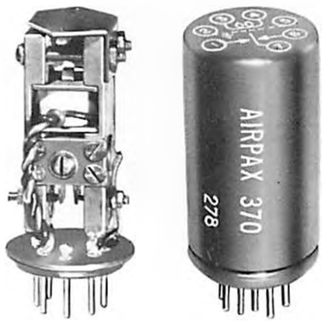


FIGURE 17. *Balanced armature type chopper operates satisfactorily while undergoing high frequency vibration.*

The type of mounting shown in Figure 18 may be helpful in permitting maximum shock and vibration resistance. The vibration resistant chopper in this mounting is Airpax Type 371. It is also available in the mountings shown in Figure 16. Also for vibration resistance, a popular mounting is the wrap around bracket. Airpax Type 372 chopper is supplied with this mounting.

Automatic direction finding equipment, used in the VHF portion of the spectrum, customarily employs low frequency modulation of the RF carrier. In subsequent demodulation the phase relationships present in the receiving loop antenna are present in the low frequency signal, and can be used to automatically orient the loop.

FIGURE 18. *Type 371 uses special bracket to balance external vibrational forces against the mount.*

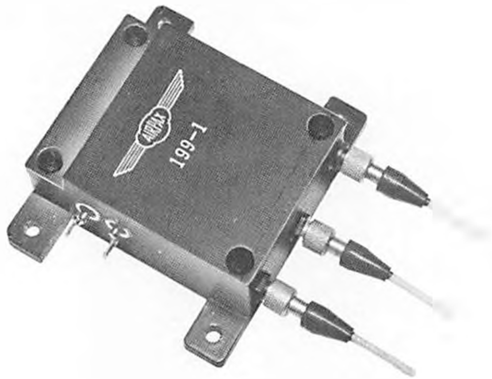


FIGURE 19. *Coaxial chopper with DPDT contacts operating at 100 CPS, is used in VHF direction finding systems in the 100-400 MC band.*

Figure 19 shows a coaxial chopper, Airpax Type 199, operating in the 100 to 400 megacycle range. This is a balanced armature structure designed for 100 cycle operation. Contacts are DPDT, break-before-make, with a dwell time of 160° and a phase angle of 30°. The standing wave ratio is better than 1.2 over the specified operating range. Connection to the chopper is made with miniature coax connectors. This chopper has been used to modulate a VHF signal for VHF direction finding. Type 199 normally operates over the range of 95 to 105 cycles, it can readily be supplied for 60 cycle operation.

Multiple pole operation is readily possible by using duplicate choppers. Choppers selected at random will have a high order of uniformity, and phase angles and dwell times can be expected to track very closely together, even when changes are forced by wide variations in ambient temperature or drive voltage. If changes occur the choppers will stay

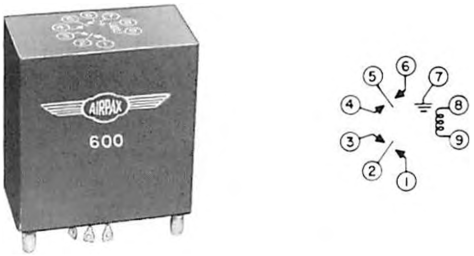


FIGURE 20. *DPDT choppers with 60 or 400 CPS drives permit complete reed isolation by using matched units.*

close together, assuming contact operating conditions are reasonably similar. As can be expected, this is even more true if units are matched (by adjustment). Somewhat surprisingly at first glance, multiple electrical operation can be as accurate as a mechanical linkage. Type 600 is a 400 cycle version. Type 800 is designed for 60 cycle operation. Complete resistive and capacitive isolation is provided with synchronous performance and common coil connections. Figure 20 shows a Type 600 chopper with pin connections.

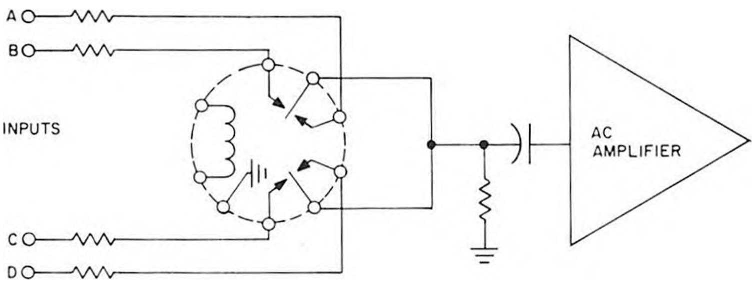


FIGURE 21. *Multiple contacts permit multiple inputs.*

Since chopper amplifiers can use SPDT constructed choppers for both modulation and demodulation, the need for multiple pole operation might be questioned. Full wave demodulation does of course permit higher output. However, a more likely requirement is input-output isolation, or mixing of many signals, Figure 21, or freedom from polarity limitations (with

full wave rectification either polarity is obtainable). If a DPDT chopper is used in a carrier type system as a reversing switch on the input as in Figure 22, improvement in circuit symmetry and a consequent improvement in common mode rejection may result.

In Figure 23, if the gain of the amplifier is sufficiently high, positive feedback may cause high frequency oscillation. Capacitive feedback may occur at the socket, across the chopper contacts or circuit wiring. Positive feedback assumes a non-inverting DC amplifier, i.e., in phase. If we add or

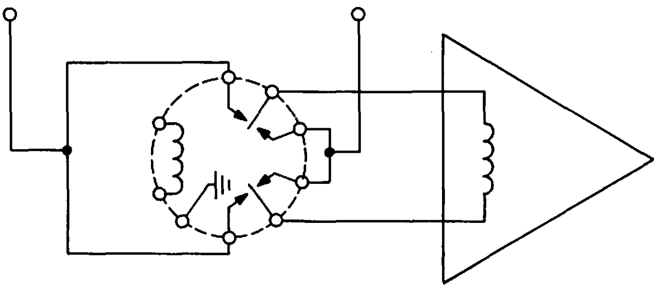


FIGURE 22. *DPDT chopper used as full wave modulator.*

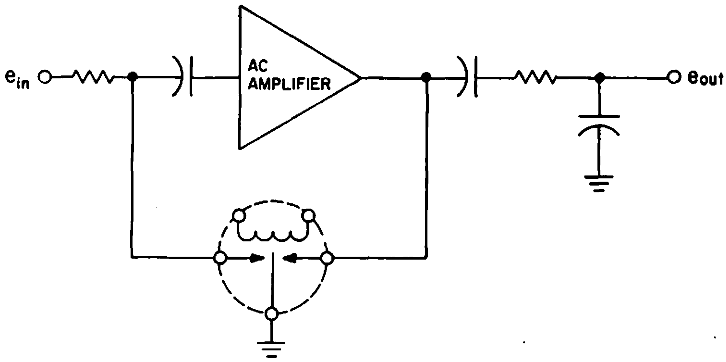


FIGURE 23. *Modulation and demodulation with a SPDT chopper.*

subtract one stage, the output polarity is inverting and the feedback will be negative. However, when used to stabilize, the stabilizing chopper amplifier must be non-inverting, i.e., (−) input delivers (−) output. The best choice, to avoid oscillation, is Model 40A, a very low capacity chopper.

Figure 24 shows one type of modulator circuit using a SPDT chopper. At high impedances, noise may be picked up inside the chopper or from adjacent wiring. If the input resistance is high, considerable noise may appear during the chopper off time.

The circuit changes occurring during the relatively short contact transit time (off time) lead some engineers to specify make-before-break. This demands perfection from the chopper contacts and is not good design practice. If any small discontinuity will disable the chopper system, it is obvious the life and reliability are thereby radically limited. In spite of

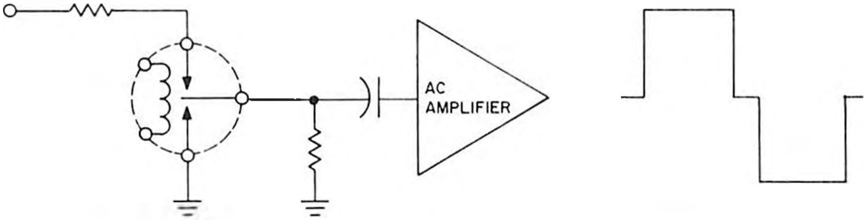


FIGURE 24. *A full wave modulation system.*

this, many systems dependent on MBB action have been built, which speaks well for chopper reliability under adverse conditions.

As the users of choppers become more aware of the possibilities and the limitations of chopper applications, the use of MBB types has reduced —a desirable objective.

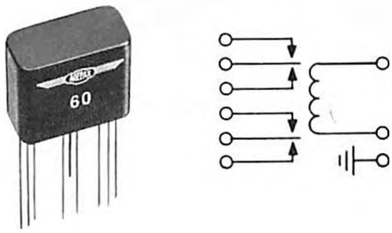


FIGURE 25. *Micro-miniature double-pole double-throw, Airpax Model 60.*

Airpax Model 60 is an example of the latest in chopper development. This is a micro-miniature size, as shown in Figure 25. It is also double-pole double-throw, and the complexity of this can be realized by the fact that it is difficult to bring so many wires through such a tiny header. The moving poles are, in addition, completely insulated from each other. Finally, this is also a low noise chopper, like Models 30A and 40A.

The double-pole configuration is most often used in chopper amplifiers of the high gain, limited bandwidth type, sometimes using transformer input and output, with one set of contacts modulating the input and the other demodulating the output. This and other circuits will appear in detail in subsequent chapters.

Noise in choppers, which we can define as that remaining when the input reduces to zero, often is demodulated and presents itself as unwanted DC offset. Quite obviously this limits the system or amplifier sensitivity.

Choppers of the kind illustrated by Types 175, 300 or 370, have noise levels in the vicinity of 50 to 100 microvolts rms, when examined at a one megohm circuit impedance and with a wide band amplifier. This is still better than can be reached with other techniques, some of which approach this low noise level, but not at a high impedance.

Airpax Models 36A and 46A are illustrated in Figures 26 and 27. In these units the residual noise is somewhere around 1 or 2 microvolts at high impedances, and it has become extremely difficult to decide when one can believe the test equipment. Stray noise from the drive circuit, if any exists, may be masked by the white noise generated by input resistors and by the input semiconductor or tube. The effective capacity between chopper coil and contacts is actually much less than 0.1 picofarad—someone has said, not even a pouf!

FIGURE 26. *Models 36A and 46A, very low noise choppers.*

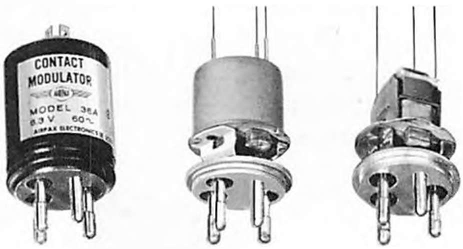


FIGURE 27. *Model 36A complete, can off, inner shields off.*

Model 36A is adjusted for optimum operation in the vicinity of 60 cycles, Model 46A for the vicinity of 400. Both have equivalent performance. It is important that extreme care be used in making electrical connections to these choppers to avoid external contact potentials, that the drive coil leads be twisted, well shielded and isolated and that the input transformer, where used, be well shielded and balanced. If connections are soldered special low contact potential solder is useful.

Figure 27 shows the internal assembly of Model 36A. This is the identical mechanism used to make Models 30A and 40A, except that the greater space permits double shielding and better drive lead isolation. The design of this unit is such that almost zero magnetic or static coupling exists between drive coil and contacts.

As of today, Airpax Model 36A can fairly be said to be the lowest-noise chopper in the world. Likewise, as of today, Airpax Model 40A is the smallest successful chopper in the world. Both are high performance

units, and will meet the full range of severe military environment for all conditions, as temperature, vibration, humidity, shock and corrosion, and at the same time will provide long life and reliability.

For operation direct from battery sources, such as 24 volts DC, Figure 28 shows Type 220, a transistorized inverter designed to drive one or two choppers. Type 220 supplies 6.3 volts at 400 CPS.

FIGURE 28. *Type 220, transistorized inverter, operates from a 24 to 26.5 volt DC source to produce 6.3 volts at 400 CPS for chopper operation.*

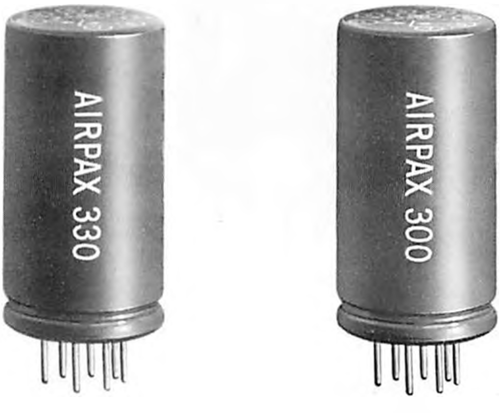


FIGURE 29. *Chopper types are available for operation from −65° to 100°C, 125°C, 150°C and 200°C.*

As illustrated in Figure 14, the chopper coil of Model 46A can be driven directly from a transistor flipflop. This is not as easily done with the larger choppers. In addition, lowest noise circuits require a drive signal balanced to ground, which is easily obtained from the internal transformer output of the Type 220 inverter.

Choppers are admirably suited to extremes of temperature. Almost any chopper will not be damaged by 100°C and will function between —65°C and 100°C. Insulation temperatures provide some limitation. The use of high temperature wire and insulating materials extends this range considerably. Thus Type 300 operates between —65°C and 100°C; Type 310 is rated up to 125°C; Type 320 up to 150°C; and Type 330, as illustrated in Figure 29, all the way up to 200°C.

There is little within the chopper likely to be contaminated by atomic radiation, unless it becomes sufficiently high to seriously injure insulation materials. Insulation in the contact circuits of most choppers is either glass or ceramic, materials relatively unaffected by radiation.

Replacement types are of many shapes, Figure 30, and sizes. Stock is carried on many part numbers, of our manufacture, and of other sources. Contact our Cambridge Division, Cambridge, Maryland, and indicate your desired replacement type.

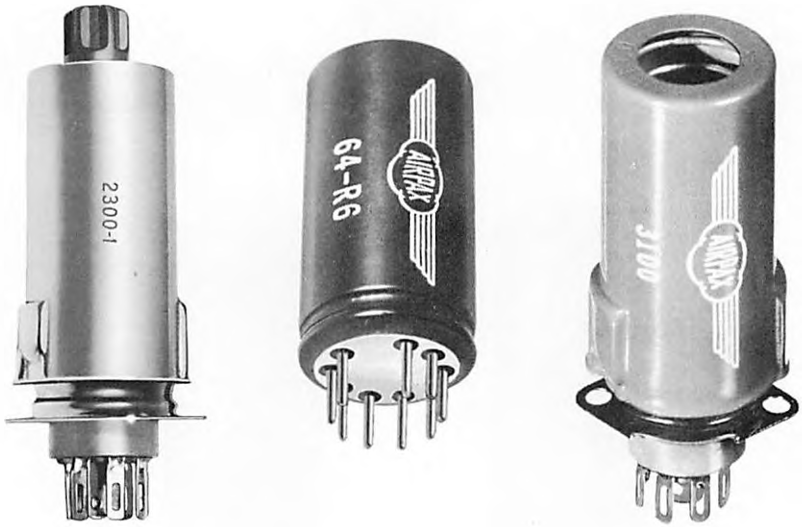


FIGURE 30. *Airpax makes replacement choppers for most types.*

SECTION III

A GLOSSARY OF CHOPPER AND AMPLIFIER TERMS

Except where pertinent,
terms general to electronics are not defined.

ASYMMETRY
: Lack of perfect balance between two halves of a square wave.

BALANCE
: The difference in dwell time, of two halves of a square wave, the difference in closure time of two contacts opposite to each other.

BBM
: Break-Before-Make.

BREAK ANGLE
: The angle of contact opening, in degrees of a driving sine wave, referred to the zero axis cross-over from which Closure Angle is measured.

CHATTER
: A discontinuity of contact closure, a break in a square wave following the original contact closing.

CHOPPER
: An electromechanical switch for the production of modified square waves of the same frequency as, and bearing a definite phase relationship to a driving sine wave.

CLOSURE ANGLE
: The angle of contact closure, measured in degrees of the sine wave, referred to the time of zero axis cross-over of a driving sine wave.

COMMON TIME
: In a MBB chopper, the angle in degrees that the contacts are all mutually closed.

CONTACT RESISTANCE
: The appearance of apparent high resistance between contacts, actually discontinuity caused by powder formation; if present to a serious degree represents chopper failure or end of life.

DEAD TIME
: Also "Off" time, in a BBM arrangement, the time neither contact is closed, the converse of Common Time.

DRIFT
: In a DC amplifier, the change in output with constant input, usually measured in terms of the DC input signal required to restore normal output; may be called out as microvolts or millivolts per hour.

DRIVE

In a chopper, the power supplied to a coil to cause action, usually measured in terms of voltage and frequency.

DWELL—DWELL TIME—ON TIME

The time in degrees during which a contact is closed, or time in milliseconds, either referred to a driving sine wave.

MBB

Make-Before-Break

MASS—MASS COMPLIANT—MASS COMPLIANT SYSTEM

The weight, the weight and spring relationship, existing in the moving structure of an electromechanical chopper.

MODIFIED SQUARE WAVE

As from a BBM chopper, with less than 180° dwell time per ½ wave.

MODULATOR

In a DC amplifier, a device for modulating DC and low frequencies to permit amplification by a carrier amplifier; usually an electromechanical chopper.

NOISE

In a chopper, that signal present across resistors between contacts and ground with zero signal supplied. In general, output other than the desired signal.

NULL

The condition of zero signal input.

NULL BALANCE

See also Offset—In a DC amplifier with zero signal input, the value of DC required at the input to return the output to zero.

OFF TIME

The time in degrees of a driving sine wave, in a BBM chopper, during which neither contact of an opposing pair is closed.

OFFSET

In a DC amplifier with zero signal input, the value, usually in microvolts or millivolts, of a DC required at the input to return the output to zero.

PHASE ANGLE

Of a chopper, the angle in degrees between a driving sine wave and the midpoint of the dwell time (the angle between the 90° point of the sine wave and the square wave center).

PHASE SHIFT

Of a chopper, a change in phase angle from any cause, measured in degrees of a driving sine wave.

PHASE BALANCE

Of a chopper, the phase angle difference between positive and negative halves of the square wave, the difference in degrees between 180° and the measured angle between square wave midpoints.

RELATIVE PHASE

Of a chopper, the correct phase relationship between a driving sine wave and the square wave, as compared to the 180° reversal of either, best measured with DC on the drive coil.

SERVO LOOP

In a servo amplifier, the entire closed loop formed by feedback from output to input. In a position servo, output position is compared to a command signal at the input.

SPIKE NOISE

In a chopper, the static field noise caused by insulating material; can be observed if the chopper is followed by a wide band amplifier.

SYMMETRY

See Balance.

ZERO SHIFT

See Offset.

SECTION IV

DEFINITIONS

The necessity of definition prior to measurement is obvious. It may be less obvious that definition is necessary to even decide on the use of choppers as compared to other methods. A chopper failure due to a sudden open connection, or broken part, is extremely rare, and since Airpax conservatively rates choppers for 10,000 hours of life (actually over 25,000 hours is not unusual), some standard of performance becomes necessary to judge the end point of life. Equipment like that shown in Figure 31 can be

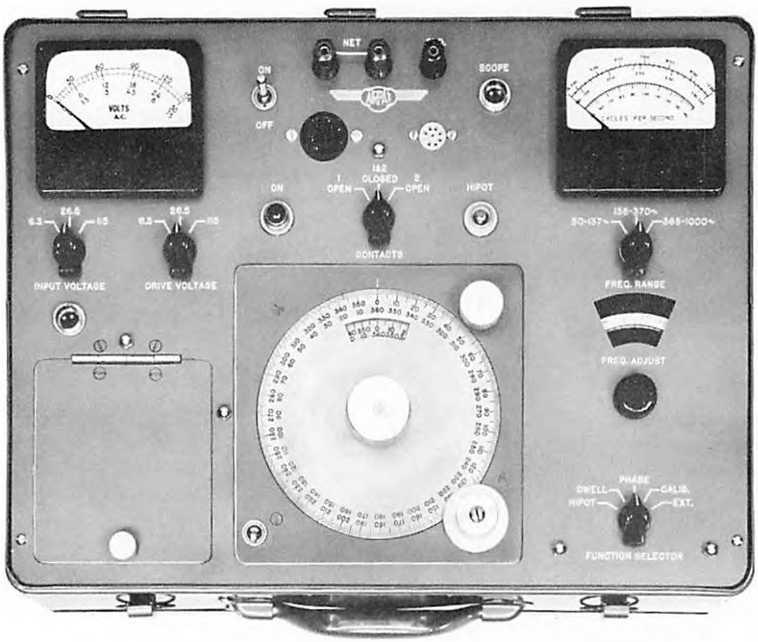


FIGURE 31. *Airpax Model TE-25 test set provides measurement of nearly all chopper parameters.*

used to measure many of the parameters defined in the following pages. Let's first tighten up our definition of "chopper".

> A chopper is an electromechanical switch for the production of modified square waves of the same frequency as, and bearing a definite phase relationship to a driving sine wave.

This covers 99% of the devices which are called choppers. We might further restrict the subject, and confine our attention to devices handling rather limited power levels.

For evaluation purposes, the chopper will be considered to be driven by a perfect sine wave which is repetitive in nature. Because, by design, the contacts maintain a constant time relationship to the driving waveform, it is possible to completely describe the operation of a chopper by investigating one cycle. For convenience, we will speak of portions of this cycle in electrical degrees, 360 degrees constituting one complete cycle of operation. The output signal, for purposes of measurement, is defined as the voltage developed across resistors connected to the contacts and supplied with DC as in Figure 32.

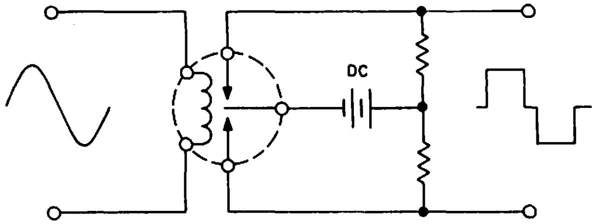


FIGURE 32. *Contact action measurement circuit.*

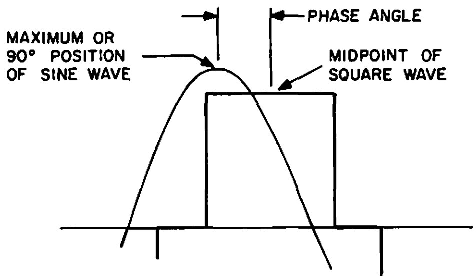


FIGURE 33. *Phase angle is measured to midpoint of square wave.*

TIME, as expressed in electrical degrees of the exciting frequency, is used as the basis of most expressions. Consider one complete cycle of a perfect sine wave used to drive the chopper, and reference all contact action to that cycle.

PHASE ANGLE, the inherent phase lag, is defined as the angle existing between the peak of the driving sine voltage and the midpoint between contact make and contact break, expressed in degrees of the driving wave, as in Figure 33. That is, phase angle is measured from the 90° (or 270°) point of the driving sine wave to the midpoint of the on-time or period of closure. In a Type 300 Airpax chopper at 400 cycles, this lag is roughly half electrical (the coil L/R relationship), and about half electromechanical.

The drawing, Figure 33, illustrates the nominal phase relationship in a Type 300 chopper at 400 cycles; for the sake of simplicity, the opposite half cycle is not shown.

DWELL TIME, also called on-time, etc., is the number of degrees each contact is closed, expressed in relation to a driving sine wave, and is illustrated in Figure 34. Obviously, it can be and sometimes is expressed

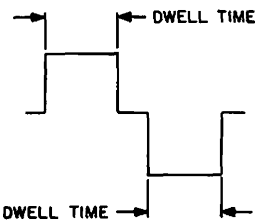


FIGURE 34. *Dwell time is the period of contact closure.*

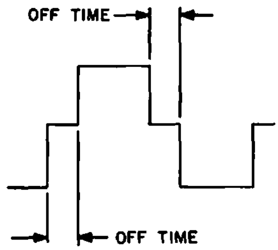


FIGURE 35. *Off time of a BBM chopper is the period when both contacts are open.*

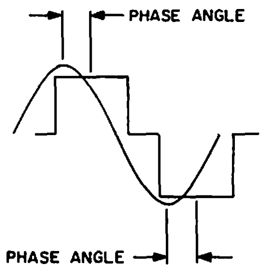


FIGURE 36. *A phase difference may exist between opposite wave halves.*

in milliseconds. This is inconvenient for measurement unless the chopper is used only at one frequency. At 400 cycles, a Type 300 chopper has a nominal dwell time of 1.05 milliseconds; i.e., 147/360° x 2.5 ms.

BALANCE OR SYMMETRY. This is the difference between positive and negative dwell times; thus, if one half were 145° and the other 140°, the balance would be 5°. It is usually specified as a maximum (actually a maximum unbalance).

OFF TIME, transit time, or dead time. The period, in degrees of the driving sine wave, during which neither contact of a BBM (break-before-make) chopper is closed. See Figure 35. It is about 0.23 ms for a Type 300 chopper at 400 cycles, and would usually be expressed as 33°; like dwell time, it occurs twice each cycle.

PHASE BALANCE refers to the possibility that the dwell times may not be perfectly symmetrical, and this might occur if the balance is in serious error. Figure 36 will clarify this. Usually this does not become a problem and choppers have phase balance within a degree or two even if the dwell time is seriously unbalanced.

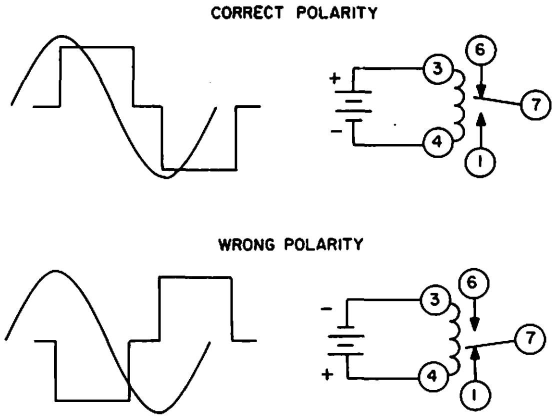


FIGURE 37. *Relative phase (polarity) of a Type 300 chopper.*

RELATIVE PHASE, or polarity. If one of many chopper polarities, such as coil leads, is reversed, the chopper may appear 180° reversed. This would prevent many servo circuits from functioning. This is illustrated in Figure 37. Polarity is easiest to test and pretty definite if it is specified in terms of DC on the coil. A chopper is of course, a polarized relay, thus

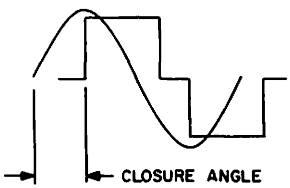


FIGURE 38. *Contact closing time is called closure angle.*

DC in one direction through the coil will always close the same pair of contacts (if they happen to be open). When the sinusoidal drive voltage is interrupted, an Airpax chopper armature comes to rest with the moving contact against either one of the fixed contacts.

CLOSURE ANGLE is the angle between the sine wave and the beginning of dwell time, as measured on the base line. See Figure 38. This angle is not often used but is of importance in some circuits.

BREAK ANGLE. This follows from the definition of closure angle. It should be noted, if used, that this angle is governed by the dwell time.

COMMON TIME. In a make-before-break (MBB) chopper connected as shown in Figure 39, the output square wave pattern will be identical to the BBM type. In this figure, when all contacts are mutually closed, the algebraic sum of the voltages across the resistors is zero, identical with the condition when all contacts are open. Common time measurement provides a more accurate and easily measured control of balance. It is the converse of off time, and dwell time is actually the square wave length plus the common time.

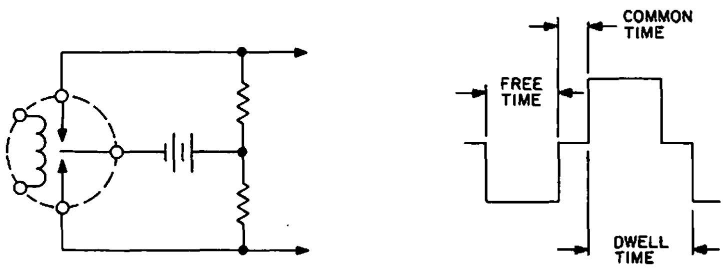


FIGURE 39. *In a make-before-break (MBB) chopper the sum of the common time plus free time is dwell time.*

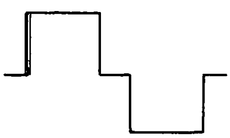


FIGURE 40. *Chatter, exaggerated in the above figure is seldom observed in Airpax choppers.*

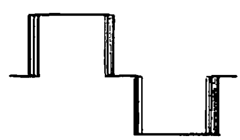


FIGURE 41. *Contact derangement is caused by vibration.*

FREE TIME is a term used to describe the square wave length of an MBB chopper connected as in Figure 39. Free time is that portion of dwell time when one contact is open; i.e., the operating contact is free of the opposite contact.

CHATTER, or contact bounce, Figure 40, usually appears quite close to the start or finish of dwell time; in fact, if it appears near the middle, it would indicate something wrong with the chopper. Chatter is rarely seen in an Airpax chopper; when it does appear, it will be found to occupy only a few degrees and is generally harmless. Usually it will appear only at a fixed frequency or a small range of frequencies.

CONTACT DERANGEMENT refers to the phase modulation caused by mechanical vibration. It is defined, per Figure 41, as the aggregate of chatter, phase modulation and unbalance caused by vibration and is measured in degrees. Any Airpax chopper withstands vibration without damage; several types are designed to perform normally under extremes of vibration.

NOISE is the undesired signal appearing between contacts and ground across the load resistor. It does not necessarily result in offset of a DC system, for one reason or another, as will be detailed in later sections. It is measured in rms, or peak, or peak-to-peak values. See Figure 42.

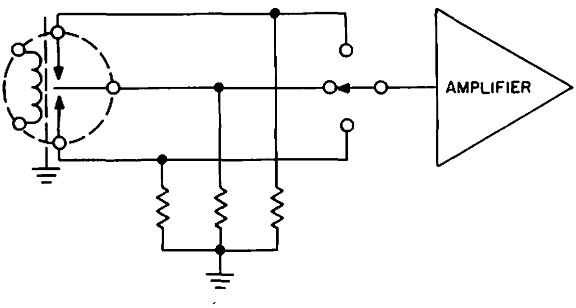


FIGURE 42. *Noise is the undesired signal from the contacts.*

Noise is not specified properly until we have defined the circuit, the ground connections, specified the amplifier frequency range, defined the value as peak, average, etc., specified the meter used, which may be peak reading, average, or rms, and until we have called out the load resistance values. Finally, noise is still not specified until we can measure the noise originating in the chopper itself, and not in connecting leads or sockets or in nearby amplifiers.

Noise is often confused with contact resistance, which is described later. Since noise is correctly defined only in the *absence* of an operating signal (i.e., it is that signal which remains when the input is removed), the difference is clear. Contact resistance, of course, may result in a random change of gain, since it alters the contact dwell time. If this appears as a fluctuating output it may be erroneously described as "noise".

OFFSET may be caused by the chopper, but is a function of the amplifier using the chopper. It is usually measured in terms of the amount of DC required at the amplifier input to return the amplifier output to zero. Offset is shown in the transfer function of Figure 43. Noise does not necessarily cause a zero offset, in fact it might reduce the offset.

CONTACT RESISTANCE is caused by the appearance of powder between contacts, usually only when operated in a "dry", i.e., very low level circuit. When viewed on an oscilloscope, the pattern fluctuates

erratically as the apparent contact resistance changes. The resistance is more likely to be a series of open circuits of very short time duration, similar to Figure 44.

It is difficult to state whether or not a "resistance" actually appears. If the oscilloscope used to view the effect has a sufficiently fast response, it may be observed that the trace approaches the base line in a very random fashion. Probably contact resistance is an open circuit whose time interval lies in the micro-microsecond region.

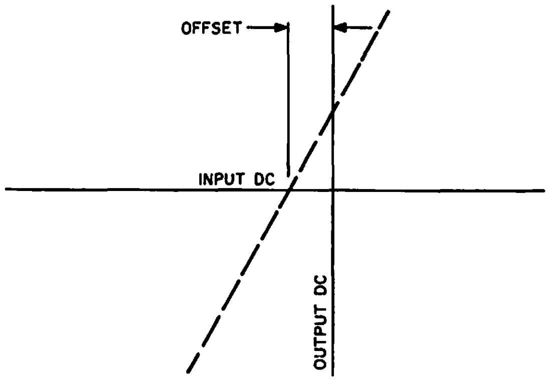


FIGURE 43. *Offset is measured as the DC input required to "zero" the amplifier output.*

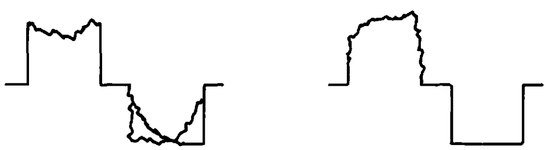


FIGURE 44. *Contact resistance shows as an erratic dwell pattern.*

LIFE. This one important factor often determines the use of a chopper. Choppers are sometimes regarded as unreliable, perhaps apt to quit working without warning. Such is absolutely not true, at least with Airpax choppers. Springs do break and coils do open up, but that kind of failure is so remote as to approach zero quantity. In the absence of other limits, a definition of the end of life is the time at which specification limits, as of phase or dwell time, are reached. As might be expected, the circuit used is important.

Contact resistance is by far the most frequent cause of failure in "dry" circuits. When appreciable currents of a milliampere or more appear

continuously, or randomly, it appears to burn or vaporize the powder, and contact resistance usually does not appear. This is why choppers often work well in circuits subject to periodic high current, as in null seeking servos. Continuous high current however, introduces an electrical erosion which may in time erode contact surfaces, with a consequent gradual reduction of dwell time and a slow increase of phase angle. The "dry" or very low current circuit is usually the worst cause of chopper failure, and except for special applications Airpax choppers are always life tested with zero current. Periodic sampling is conducted, but at very low level to avoid concealing the contact resistance defect.

TEST VOLTAGES from contacts to ground are usually limited by an internal air gap to a maximum of 200 volts AC, but flashover causes no damage. The coil test voltages vary considerably, from 1500 volts for 100-volt coils to 200 volts on 6-volt coils. Insulation resistance is very high, being limited primarily by leakage across glass at the pins.

SECTION V
MEASUREMENT

Measurement of chopper parameters is usually done oscilloscopically. Unfortunately, most of the available phase meters depend on zero axis cross-over to measure phase. (The wave to be measured is converted to a square wave.) The comparison of sine wave to step wave leaves these meters somewhat frustrated.

There are three possible means of measurement. We can display the pattern in rectilinear fashion and measure lengths with a ruler as in Figure 45; arrange the pattern in a circle, Figure 46; or mix a marker pip with the signal, Figure 47.

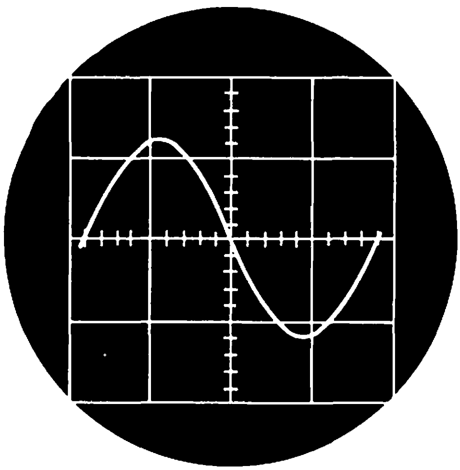


FIGURE 45. *Trace is several degrees in width.*

The rectilinear display of Figure 45, is the least desirable. The accuracy of the method obviously depends upon the linearity of the sweep circuit employed as well as the measurement of lengths.

Assuming a 5-inch scope tube is used with a trace 4 inches long, then 1 degree equals $4''/360^\circ$ or .011 inches. If an accuracy in the order of 1 degree is to be obtained, it is necessary to make a linear measurement on the face of the scope tube accurate to .01 inch.

This is rather impractical if the curvature of the tube face and parallax error are to be considered. Also most sweep circuits employed in oscilloscopes are not completely linear. Therefore if an oscilloscope is used, it is essential first, that the representation of 1 or 2 degrees be of a sufficient dimension to allow accurate measurement, and second, that the sweep be linear.

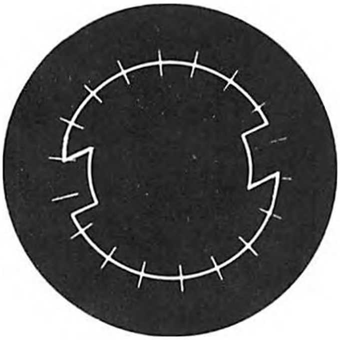


FIGURE 46. *Polar presentation provides a longer scale.*

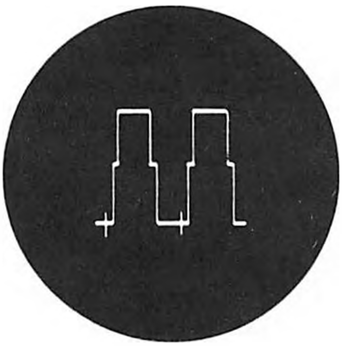


FIGURE 47. *Marker pips permit measurement of chopper parameters.*

Polar presentations provide a reading length π times longer, and also permit a 180° phase reversal to minimize error due to distortion and imperfect deflection accuracy. If a circle of 5-inch diameter is used, a trace length of 15.7 inches represents 360 degrees so that 1 degree equals 15.7/360° or .044 inch. This represents a minimum distance which can be practically read on a scope tube face. The linearity requirement can be readily fulfilled by using low distortion amplifiers, and obtaining an accurate 90-degree displacement.

No difficulty will be experienced in obtaining two voltages 90 degrees out of phase at a fixed frequency, and special circuits will yield the desired results over the expected frequency range of the chopper under test. The major errors occur in the CRT. The deflection plates are seldom exactly at 90 degrees to each other, and the circle becomes elliptical due to deflection distortion. Errors due to distortion can be practically eliminated by selection of the cathode ray tubes used in the measuring equipment.

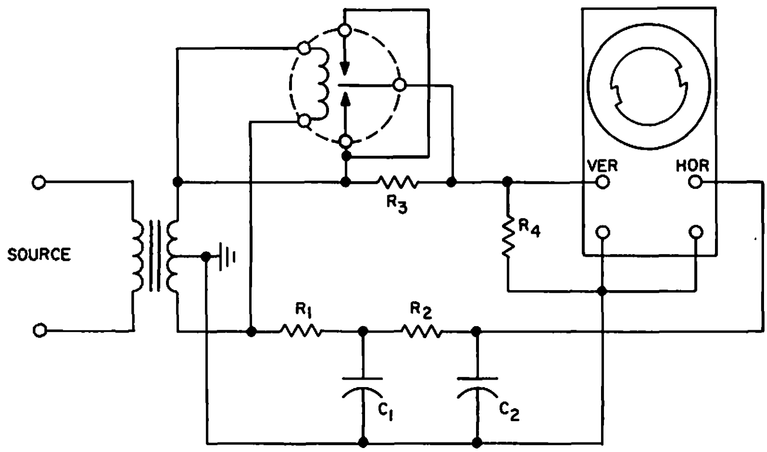


FIGURE 48. *Chopper contact performance is easily displayed in polar fashion.*

After the circle has been established, it is simply necessary to drive the chopper from one of the two signals applied to the CRT and to interrupt the circle during the off time of the chopper contacts. The resulting pattern is shown in Figure 46. The foregoing method of dwell time measurement is the result of an extensive study to determine the most feasible system consistent with production practice.

A relatively simple circuit for a fixed frequency is shown in Figure 48. To measure dwell time, it only remains to place an azimuth dial over the pattern and to read this time directly in degrees. Some error will of course remain, and can be further minimized. Referring again to Figure 46, if the two dwell periods are interchanged in time, errors in measurement can be directly observed. This can be done by reversing the polarity of the drive coil. Errors of measurement being reasonably the same on the same side of the scope, the measuring accuracy is improved.

In setting up the circuit of Figure 48, first, obviously, the source must deliver a low distortion sine wave of sufficient power and voltage to operate the chopper. R_1, C_1, R_2, C_2 must be adjusted to 90° phase shift

at the desired frequency, and this is most easily found by shorting out the chopper contacts and adjusting to a circle. If there is no distortion the 90° position can be observed within one or two degrees. R_1 is a fairly low resistance of a few thousand ohms, whose main use is to keep the circuit

FIGURE 49. *Airpax production test equipment uses polar presentation.*

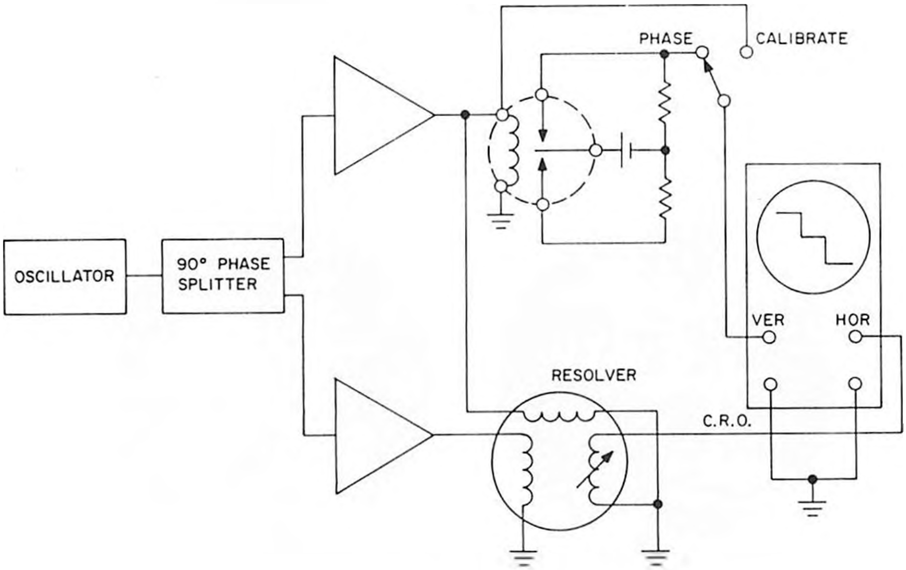


FIGURE 50. *A resolver simplifies measurement of phase angle.*

impedance low. R_3 is a high resistance of perhaps 100,000 ohms and serves to make the contact switching point more readable.

Figure 49, depicts a rather elaborate piece of chopper test equipment, in which special effort provides low distortion amplifiers, push-pull deflection and other methods of obtaining a high order of accuracy. Signals

having a 90° relationship are obtained with half-lattice networks as in single sideband transmission, and operation over a wide frequency range is therefore practical and easy.

Phase angle is most easily read by the use of a resolver driven from a calibrated dial. In Figure 50, the output of a two-phase resolver is

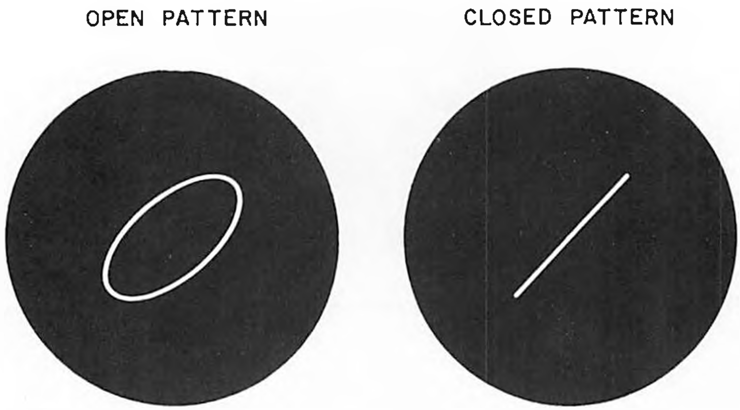


FIGURE 51. *Resolver position giving straight line is termed "zero".*

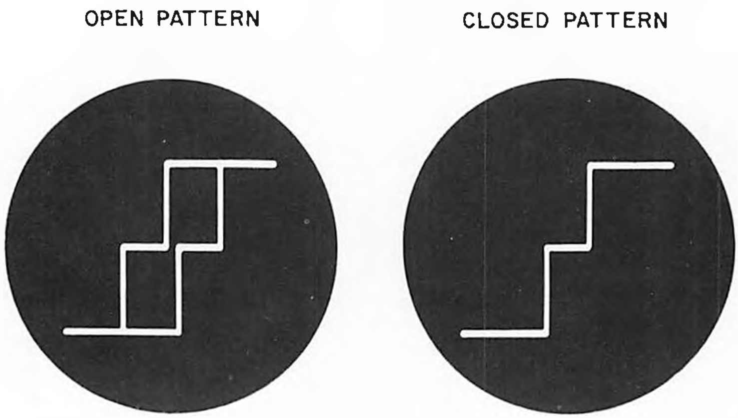


FIGURE 52. *Difference angle to close the step is chopper phase.*

compared first against the chopper drive and then to the signal from the contacts, the difference angle being read on a dial. This point of closure of the pattern appears on the right in Figures 51 and 52.

The angular difference of the resolver shaft is the phase angle of the chopper. The accuracy of this method depends upon the accuracy of the

resolver, low distortion waveforms, 90 degree displacement between the two voltages and low phase shift in the scope amplifiers. All of these factors are controllable and hence, the accuracy of this method can be very good.

Choppers from Airpax receive several complete tests before shipment, and all are given 50 hours of operation. (We thereby raise our quality level, eliminating any early failures.) Results indicate this to be successful and our choppers exhibit a very high order of reliability and uniformity.

The following pictures show the manufacturing, adjustment and test equipment used at the Airpax Cambridge Division plant. Each chopper is put through exhaustive tests before final acceptance.

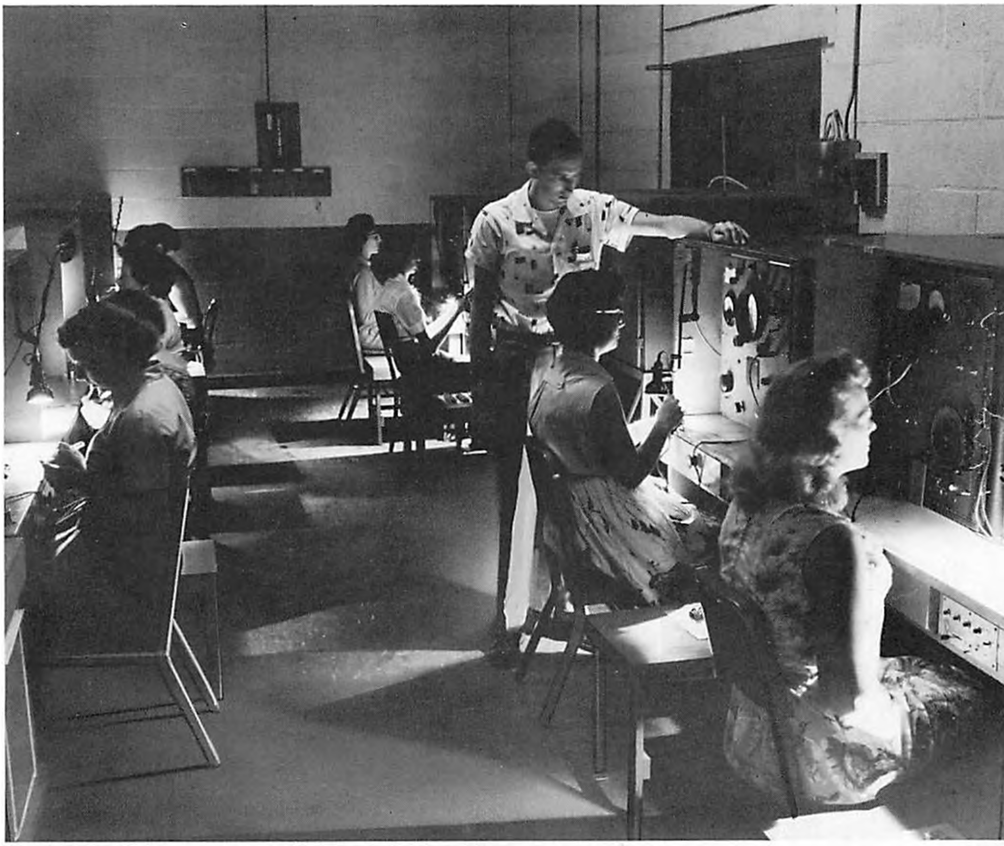


FIGURE 53. *A view of the "set-up" room where adjustments of phase and dwell time are made on choppers from production.*

FIGURE 54. *Oscilloscopic patterns permit precise adjustments.*

FIGURE 55. *One of many final test positions. Tests include insulation resistance, noise and hi-pot on finished choppers.*

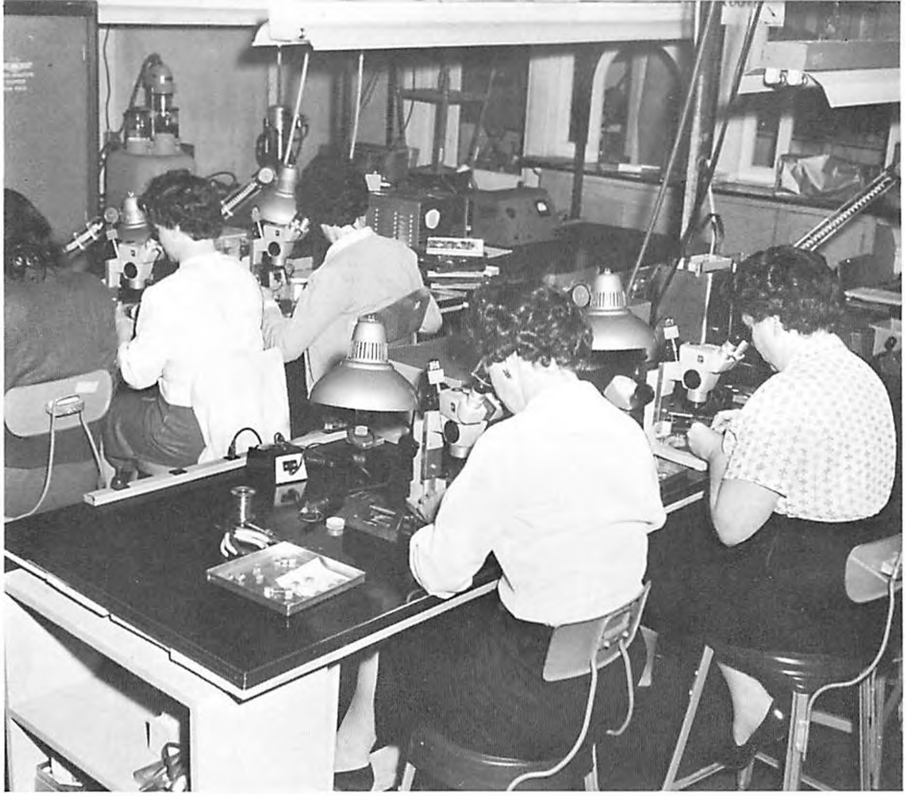
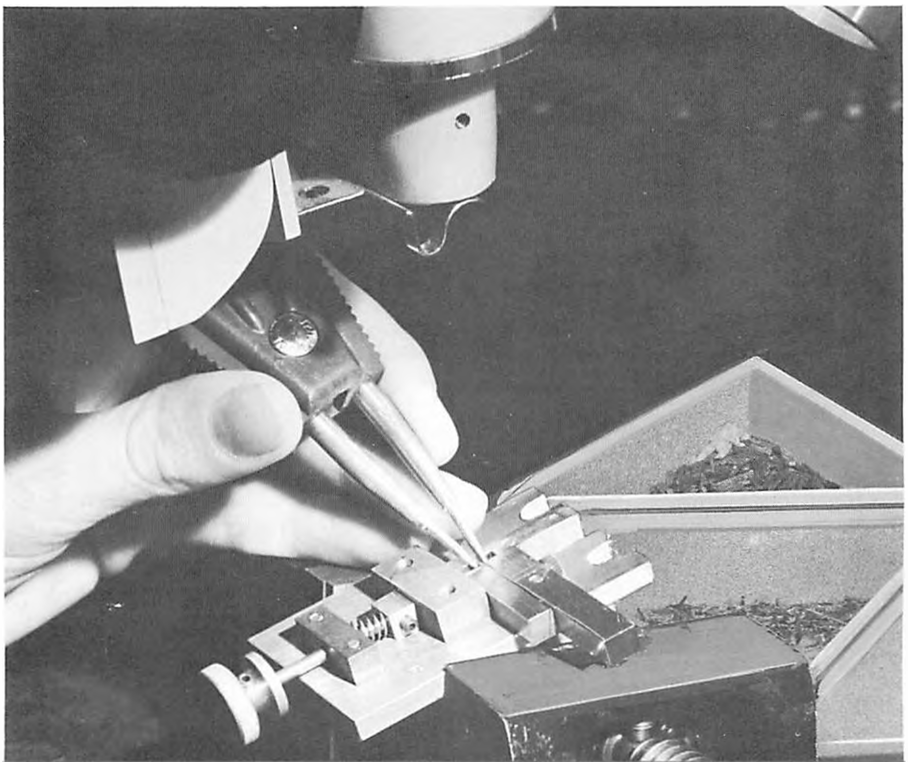


FIGURE 56. *Micro-midget chopper assembly area; precise adjustment is assured through the use of binocular microscopes.*

FIGURE 57. *Model 40A being worked on under microscope.*



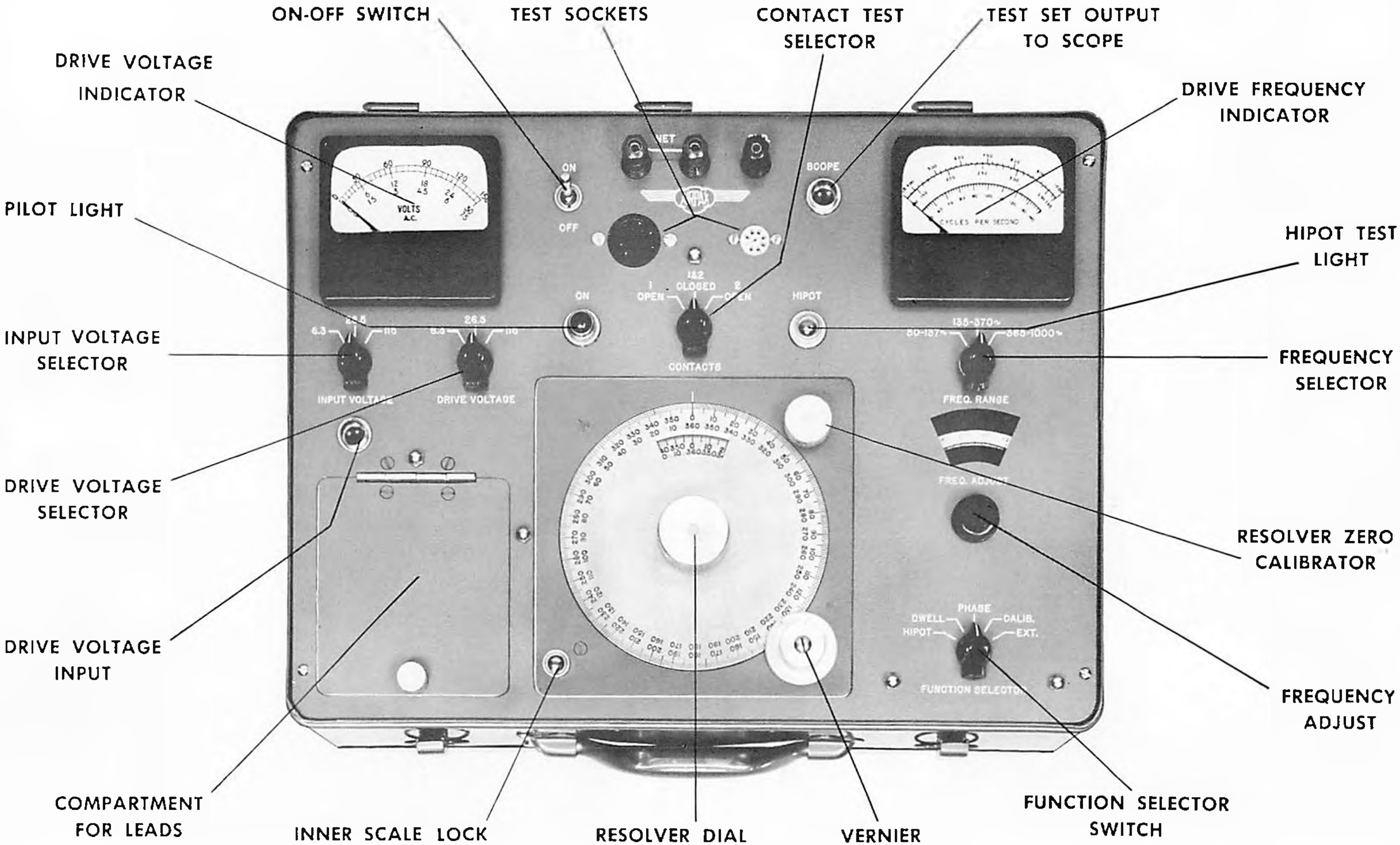


FIGURE 58. *Portable test set for chopper measurement.*

SECTION VI
MEASUREMENT EQUIPMENT

The principle of having available a continuously variable phase permits the third type of presentation mentioned earlier, in which a marker pip is used to measure dwell time. A portable test set, Airpax TE-25, is shown in Figure 58, which gives a pattern previously shown in Figure 47. An external source of signal is needed to drive the chopper, and an oscilloscope to view the output. The equipment has a self-contained time base which connects to the horizontal amplifier. It is usable on any chopper to measure phase, dwell time, chatter and balance, and has a self-contained hi-pot test. The drive may be any voltage up to 150, or any frequency between 50 and 1000 cycles. The measurement accuracy is about 1 degree.

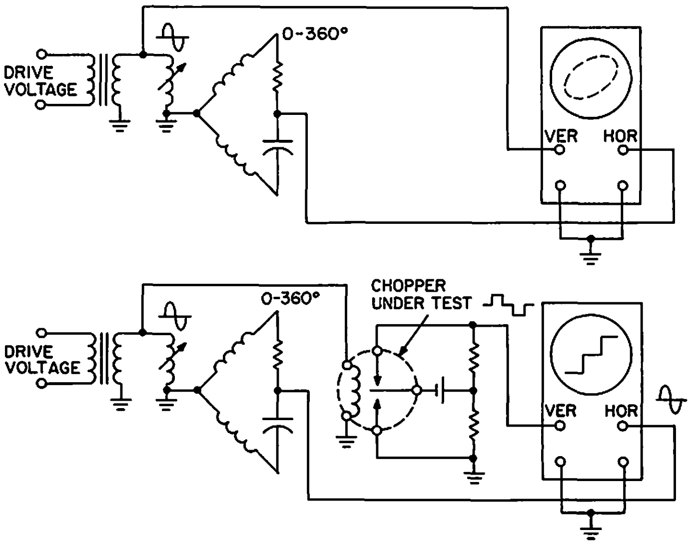


FIGURE 59. *Simplified phase measurement circuit.*

Figure 59, describes the basic circuits used for phase measurement. The two-phase resolver is supplied with a capacitive phase splitter, which is adjusted to 90° for each nominal frequency setting, a simple matter of setting a calibrated dial. The resolver provides highly accurate phase angles which are used for all measurements. To provide a simple method of zero adjustment, the housing of the resolver is rotated to zero, while shaft rotation supplies the measurement.

The measurement of chopper dwell time uses a very similar circuit, greatly simplifying the equipment and improving the accuracy thereby.

In Figure 60, note that a marker pip generator is driven from the resolver output. After amplification, the resolver output is clipped and limited, integrated, and used to trigger a one pulse multivibrator. This pulse is again differentiated and the positive going pulse deleted. The spike finally obtained is mixed with the signal from the chopper contacts (or any signal). It can now be moved at will along the wave being examined.

It is, of course, quite practical to adjust the pip starting position in the same manner as with a phase measurement, however, it is inconvenient and time consuming to make this exacting adjustment. Avoiding this, the main dial carries a fiduciary or auxiliary dial which is locked and unlocked by a panel switch, thus "zero" or pip start can be at any dial position.

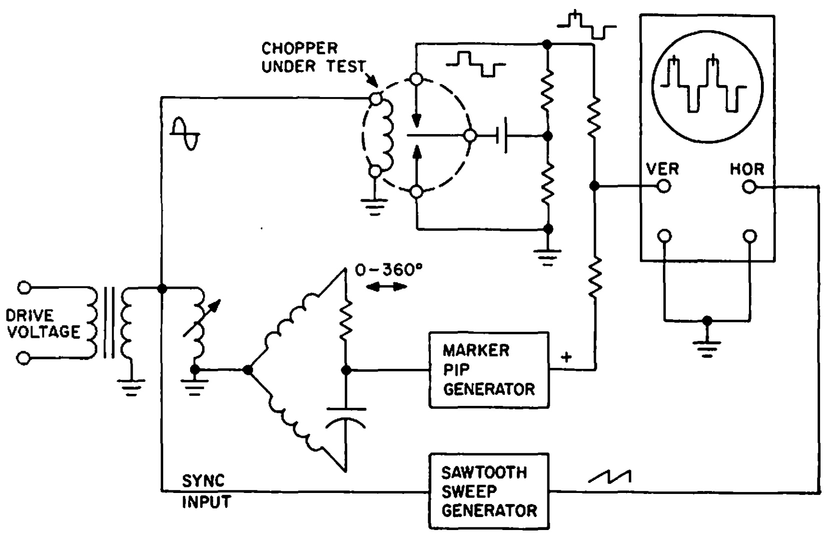


FIGURE 60. *Dwell time measurement circuit.*

To provide a way of locating low insulation resistance or shorts, a 250 volt hi-pot test is self-contained, as in Figure 61. A neon bulb is used to indicate breakdown.

Model TE-25 is particularly useful for establishing phase angles between odd wave forms lying within its range, and Figure 62 shows the marker obtainable. Terminals are also provided to connect phase shifting networks ahead of a chopper when it becomes necessary to adjust circuit or chopper phase.

The method of adjusting chopper phase angle, using a resistor or resistor-condenser combination in series with the drive voltage, was described in Section II, Figure 13. It was shown that a chopper with a nominal phase angle of 65° can be conditioned to produce other phase angles.

Life testing is undoubtedly one of the most important chopper tests. Probably the life test circuit should simulate the circuit and conditions expected. Not knowing the application, we look for the worst. The worst seems to be dry circuit conditions and room ambients (apparently "hot" operation helps Airpax choppers, certainly it does no harm). It is important that the measuring circuits do not "clean up" contacts, which will happen if the current reaches as high as one milliampere. Operation with

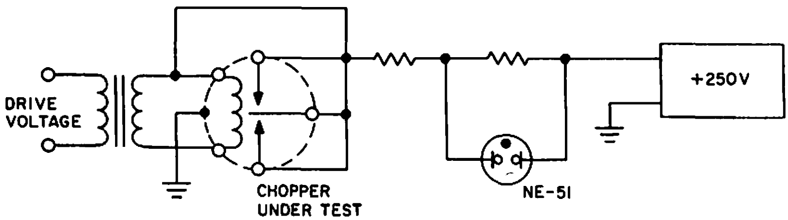


FIGURE 61. *Hi-pot test circuit.*

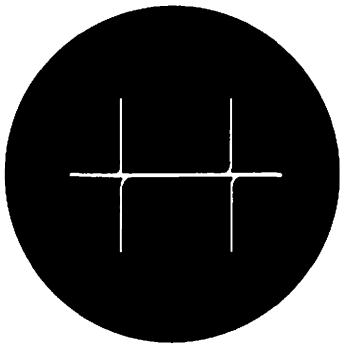


FIGURE 62. *TE-25 test set provides marker pips for various measurements.*

zero current seems to help powder formation, causing contact resistance, as mentioned earlier. Contact resistance due to powder formation produces irregular output pulses similar to a rapid series of open circuits. Irregularity is due to shifting of the powder particles as the reed vibrates. An oscilloscopic pattern of contact resistance was shown in Figure 44.

It is under these worst conditions that we find life figures greatly beyond 10,000 hours, in fact, we have observed over 25,000 hours on Airpax choppers operating at 400 cycles.

Figure 63 describes a test circuit measuring average dwell time. As any deterioration will be reflected by the contacts as a change, usually a reduction, of dwell time, this is a most sensitive measurement. Results can be easily read on a DC meter movement, recording or indicating, as it will integrate the pulse arrival to provide an average dwell time. It is also easily calibrated by shorting contacts (open the chopper drive lead).

Measurement is made at low level, 50 millivolts, 50 microamperes, to avoid a possible "cleanup" of the contacts. The 50 microampere DC meter permits visual observation if a recorder is not available. If the meter has a scale of 100, the reading will be about 80, the meter being set to full scale by shorting the contacts and adjusting the series resistor (shown as 1000 ohms). Contact resistance shows as erratic and reducing meter readings.

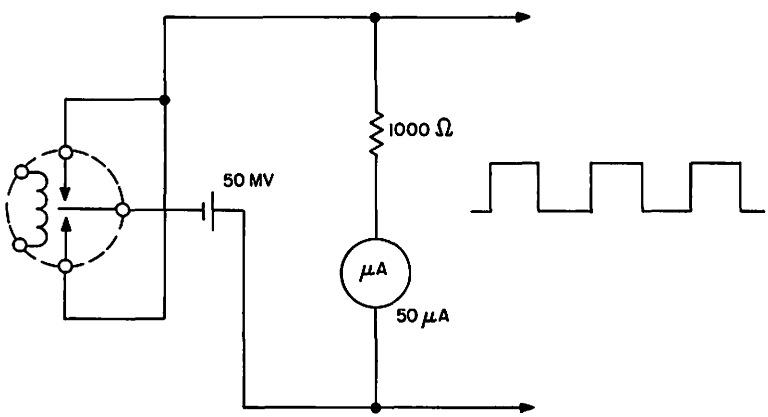


FIGURE 63. *Life test measurement circuit avoids cleaning up contacts.*

A standard D'Arsonval movement should be used, of course, since such an instrument will integrate the pulses delivered by the chopper contacts with fair accuracy.

Readings should be taken frequently when life testing choppers, and we firmly recommend not more than 24-hour periods. It is possible for contact resistance to appear, then disappear. Later tests may show an apparently good chopper but there may have been an open contact condition in the meantime. Figure 64 is automatic life test equipment handling several hundred choppers. Once every few hours the dwell time is measured and recorded on strip charts. Complete records are kept and life test information is available for many years of testing. Once every hundred hours an operator examines all parameters of each chopper such as phase angle, noise, dwell time, hi-pot, etc., which is kept with the recorded data.

FIGURE 64. *Double rack life test gear.*

Results of such records are illustrated in Figure 65, which shows the average dwell time of a normal chopper (top), and of a chopper which has developed excessive contact resistance.

Contacts which function best in dry circuits are usually not the best for relatively heavy currents. The effect of several milliamperes of current on a standard Airpax chopper is to slowly wear away the contact surface, reducing the dwell time noticeably. A Type 300 chopper carrying 2 ma DC at 100 volts, for example, will wear enough in 1000 hours to get close to the

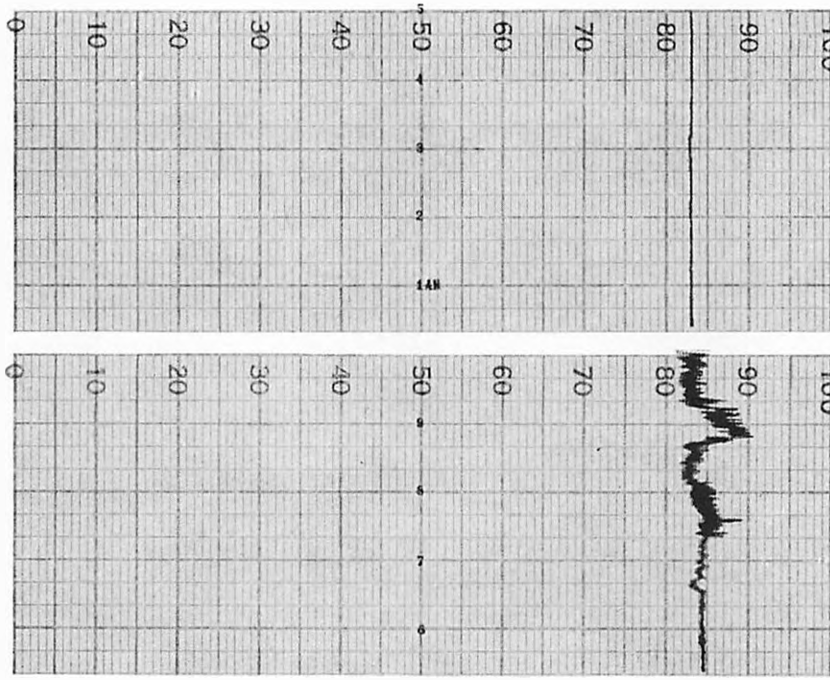


FIGURE 65. *Strip chart shows normal and erratic contact behavior.*

lower specification limit of dwell time, and upper limit of phase angle. (If you have such an application, consult our Engineering Department. We can furnish choppers handling substantial power.)

Noise has been referred to as contact resistance, previously shown in Figure 44, or as amplifier offset. By our definition, as described earlier, it is the stray or unwanted signal appearing at the contacts, when there is zero signal input. It must also be noted that the measured amount of noise may bear only a general relationship to resultant offset of an amplifier, due to the noise phase relative to signal, and other factors such as bandwidth.

A fundamental limitation[3] is set by the noise of thermal agitation, or Johnson noise, which is defined by the relationship:

$$E_{eff} = \sqrt{4KTRF}$$

where $K = 1.37 \times 10^{-23}$, T is degrees Kelvin, R the resistance involved, and F, the bandwidth.

If the resistor has a value of 1 megohm and a wide band amplifier of 100 KC is used, at room temperature (300°K), the noise level will be 45

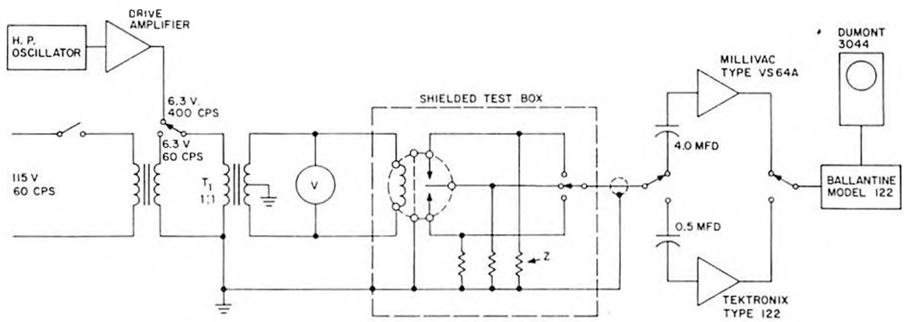


FIGURE 66. *Noise test circuit as presently used.*

FIGURE 67. *Photo of noise test equipment.*

microvolts effective value, or about 360 microvolts peak-to-peak. This applies to wire-wound resistance values, carbon resistor noise is likely to be considerably higher.

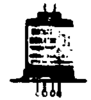
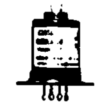
[3]DC Amplifier Stabilized for Zero and Gain, Williams, Tarpley and Clark, AIEE transactions, Vol. 67, 1948.

SECTION VII

A REPRESENTATIVE LISTING OF THE MOST COMMONLY USED AIRPAX MINIATURE AND MICRO-MIDGET ELECTROMECHANICAL CHOPPERS

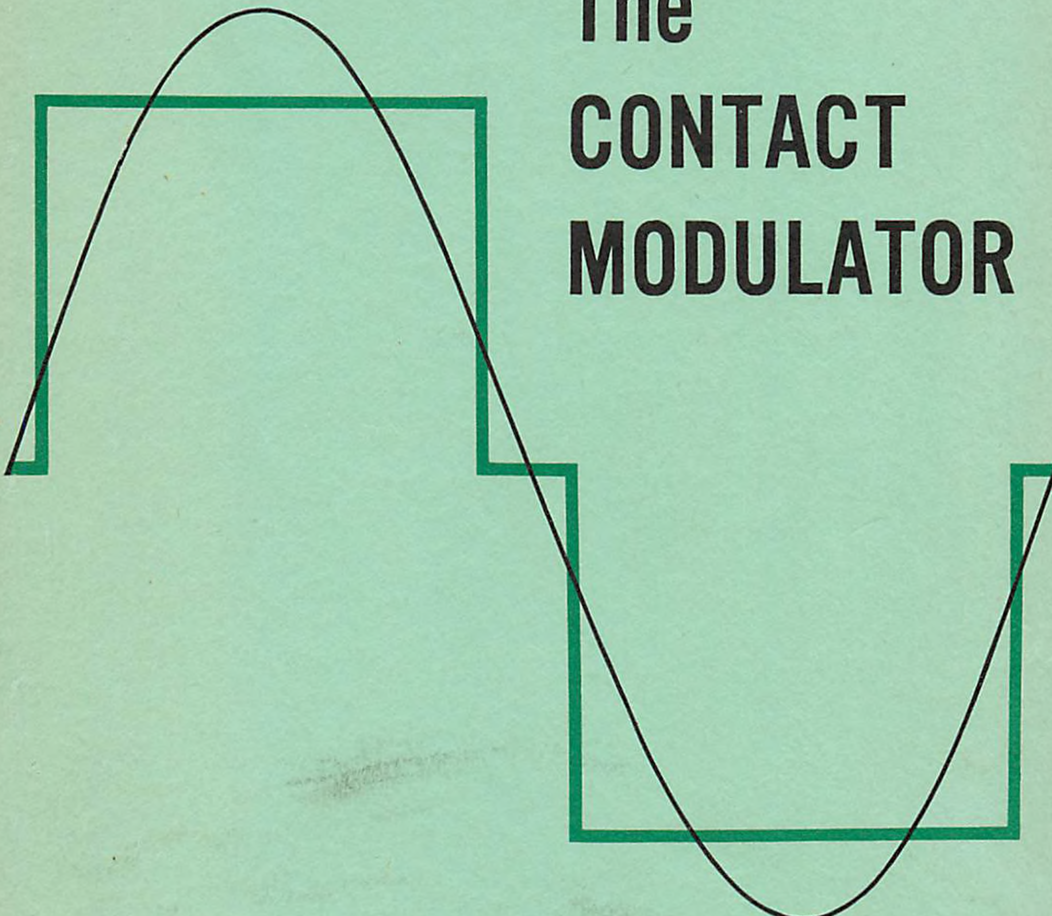
Parameter	Series 175	Series 190	Series 300	Series 310	Series 360	Series 370
						
Drive Voltage and Frequency	6.3 volts @ 60 cps	6.3 volts @ 50 cps	6.3 volts @ 400 cps	6.3 volts @ 400 cps	6.3 volts @ 400 cps	6.3 volts @ 400 cps
Noise	50 uv RMS max., 1 megohm load	50 uv RMS max., 1 megohm load	200 uv RMS max., 1 megohm load	200 uv RMS max., 1 megohm load	200 uv RMS max., 1 megohm load	200 uv RMS max., 1 megohm load
Switching	SPDT, BBM	SPDT, BBM	SPDT, BBM	SPDT, BBM	SPDT, BBM	SPDT, BBM
Shock	100G	100G	100G	100G	100G	100G
Vibration	5G, to 500 cps	5G, to 500 cps	5G, to 500 cps	5G, to 500 cps	5G, to 500 cps	15G, to 2500 cps
Weight	1½ ounces max.	1½ ounces max.	1½ ounces max.	1½ ounces max.	1½ ounces max.	1½ ounces max.
Life	5,000 hrs. min.	2,000 hrs. min.	5,000 hrs. min.	2,000 hrs. min.	5,000 hrs. min.	2,000 hrs. min.
Temp. Range	—65°C to +100°C	—65°C to +100°C	—65°C to +100°C	—65°C to +125°C	—65°C to +100°C	—65°C to +125°C
Phase Angle	20° ±5°	20° ±6°	65° ±15°	65° ±15°	55° ±15°	58° ±12°
Dwell Time	150° to 177°	155° to 177°	130° to 165°	130° to 165°	145° to 175°	145° to 175°
Dissymmetry	15° max.	15° max.	15° max.	15° max.	15° max.	15° max.
Transit Time	2°	2°	2°	2°	2°	2°
Bounce	4° max.	4° max.	4° max.	4° max.	4° max.	4° max.
Contact Rating	100 VDC, 2 MA max.	100 VDC, 2 MA max.	100 VDC, 2 MA max.	100 VDC, 2 MA max.	100 VDC, 2 MA max.	100 VDC, 2 MA max.
Coil Resistance	155 ohms	155 ohms	155 ohms	155 ohms	155 ohms	148 ohms
Coil Impedance	185 ohms	185 ohms	270 ohms	270 ohms	235 ohms	235 ohms
Airpax Spec.	#232	#234	#219	#239	#242	#243

Parameter	Series 400 	Series 600 	Series 1400 	Series 1800 	Series 2300 	Series 2400 
Drive Voltage and Frequency	6.3 volts @ 400 cps	6.3 volts @ 400 cps	6.3 volts @ 60 cps	6.3 volts @ 60 cps	6.3 volts @ 400 cps	6.3 volts @ 60 cps
Noise	200 uv RMS max., 1 megohm load	200 uv RMS max., 1 megohm load	70 uv RMS max., 1 megohm load	70 uv RMS max., 1 megohm load	4 uv RMS max., 100 ohm load	2 uv RMS max., 100 ohm load
Switching	SPDT, BBM	DPDT, BBM	SPDT, MBB	DPDT, MBB	SPDT, BBM	SPDT, BBM
Shock	100G	100G	100G	100G	100G	100G
Vibration	5G, to 500 cps	5G, to 500 cps	5G, to 500 cps	5G, to 500 cps	5G, to 500 cps	5G, to 500 cps
Weight	1½ ounces max.	2 ounces max.	1½ ounces max.	2 ounces max.	1½ ounces max.	1½ ounces max.
Life	2,000 hrs. min.	2,000 hrs. min.	2,000 hrs. min.	2,000 hrs. min.	2,000 hrs. min.	2,000 hrs. min.
Temp. Range	—65°C to +100°C	—65°C to +100°C	—40°C to +85°C	—40°C to +85°C	—65°C to +100°C	—65°C to +100°C
Phase Angle	65° ±15°	65° ±15°	21° ±10°	21° ±10°	30° ±15°	20° ±5°
Dwell Time	130° to 165°	130° to 165°	178° to 218°	180° to 220°	130° to 170°	150° to 177°
Dissymmetry	15° max.	15° max.	20° max.	20° max.	18° max.	15° max.
Transit Time	2°	2°	0°	0°	2°	2°
Bounce	4° max.	4° max.	4° max.	4° max.	4° max.	4° max.
Contact Rating	100 VDC, 2 MA max.	100 VDC, 2 MA max.	100 VDC, 2 MA max.	100 VDC, 2 MA max.	100 VDC, 2 MA max.	100 VDC, 2 MA max.
Coil Resistance	155 ohms	78 ohms	108 ohms	54 ohms	300 ohms	155 ohms
Coil Impedance	270 ohms	131 ohms	120 ohms	60 ohms	315 ohms	185 ohms
Airpax Spec.	#264	#244	#247	#252	#249	#248

Parameter	Model 30A	Model 33A	Model 36A	Model 40A	Model 43A	Model 46A
						
Drive Voltage and Frequency	6.3 volts @ 60 cps	6.3 volts @ 60 cps	6.3 volts @ 60 cps	6.3 volts @ 400 cps	6.3 volts @ 400 cps	6.3 volts @ 400 cps
Noise	6 uv RMS max. 1 megohm load	6 uv RMS max. 1 megohm load	6 uv RMS max. 1 megohm load	6 uv RMS max. 1 megohm load	6 uv RMS max. 1 megohm load	6 uv RMS max. 1 megohm load
Switching	SPDT, BBM	SPDT, BBM	SPDT, BBM	SPDT, BBM	SPDT, BBM	SPDT, BBM
Shock	100G	100G	100G	100G	100G	100G
Vibration	15G, 55-2500 cps	15G, 55-2500 cps	15G, 55-2500 cps	15G, 55-2500 cps	15G, 55-2500 cps	15G, 55-2500 cps
Volume	0.151 cu. in.	0.453 cu. in.	0.455 cu. in.	0.151 cu. in.	0.453 cu. in.	0.455 cu. in.
Weight	9 grams	22½ grams	20½ grams	9 grams	22½ grams	20½ grams
Life, min.	2,000 hrs.	2,000 hrs.	2,000 hrs.	2,000 hrs.	2,000 hrs.	2,000 hrs.
Temp. Range	−65°C to +100°C	−65°C to +100°C	−65°C to +100°C	−65°C to +100°C	−65°C to +100°C	−65°C to +100°C
Phase Angle	25° ±10°	25° ±10°	25° ±10°	65° ±15°	65° ±15°	65° ±15°
Dwell Time	155° to 185°	155° to 185°	155° to 185°	140° to 185°	140° to 185°	140° to 185°
Dissymmetry	15° max.	15° max.	15° max.	15° max.	15° max.	15° max.
Transit Time	2°	2°	2°	5°	5°	5°
Bounce	4° max.	4° max.	4° max.	4° max.	4° max.	4° max.
Contact Rating	10 VDC, 2 MA max.	10 VDC, 2 MA max.	10 VDC, 2 MA max.	10 VDC, 2 MA max.	10 VDC, 2 MA max.	10 VDC, 2 MA max.
Coil Resistance	310 ohms	310 ohms	310 ohms	85 ohms	85 ohms	85 ohms
Coil Impedance	330 ohms	330 ohms	330 ohms	115 ohms	115 ohms	115 ohms
Airpax Spec.	#266-1	#266-5	#266-7	#288-7	#288-8	#288-9

AIRPAX

The CONTACT MODULATOR



PART 2: Definitions and Measurement

THE SECOND OF A SERIES

AIRPAX

THE CONTACT MODULATOR

PART II

DEFINITIONS AND MEASUREMENT

WRITTEN BY

THE ENGINEERING STAFF

OF

AIRPAX ELECTRONICS

INCORPORATED

Cambridge, Maryland • Fort Lauderdale, Florida

H. A. COOK, President

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FOREWORD

PART II of this series of booklets on the Contact Modulator is concerned with the definition and measurement of chopper parameters.

Airpax was instrumental in coining and bringing into accepted usage many of the terms applied to chopper technology.

AIRPAX ELECTRONICS
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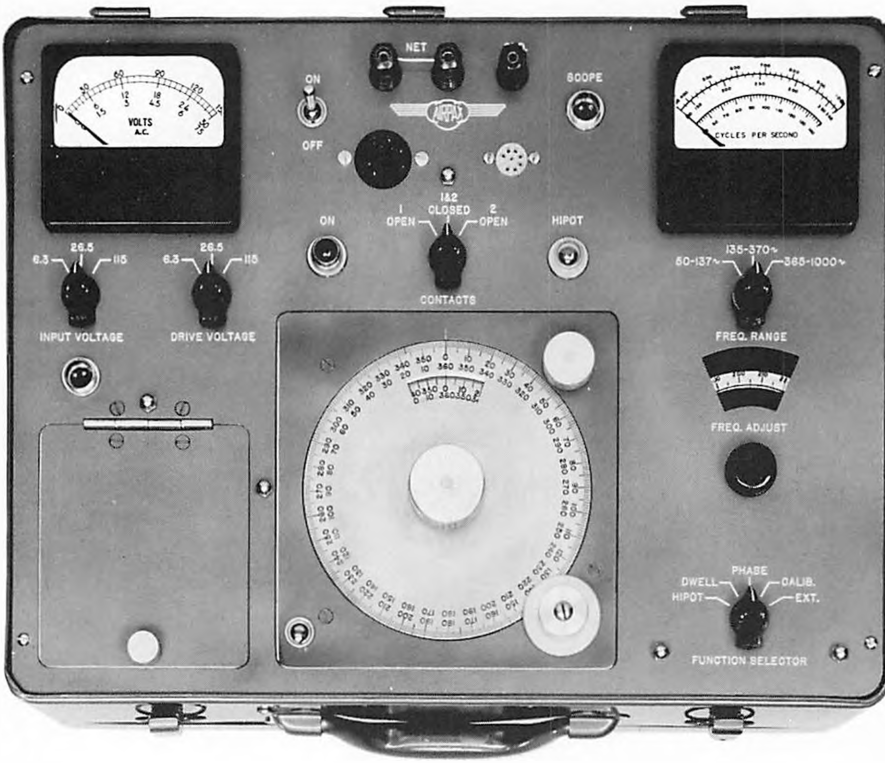


FIGURE 1. *Airpax model TE-25 portable test set provides measurement of nearly all chopper parameters.*

SECTION I

DEFINITIONS

The necessity of definition prior to measurement is obvious. It may be less obvious that definition is necessary to even decide on the use of choppers as compared to other methods. (See PART I, WHY USE CHOPPERS?). A chopper failure due to a sudden open connection, or broken part, is extremely rare, and since Airpax conservatively rates choppers for 5000 hours of life (actually over 25,000 hours is not unusual), some standard of performance becomes necessary to judge the end point of life. Equipment like that shown in Figure 1, can be used to measure many of the parameters defined in the following pages. Let's first tighten up our definition of "chopper."

> A chopper is an electro-mechanical switch for the production of modified square waves of the same frequency as, and bearing a definite phase relationship to a driving sine wave.

This covers 99% of the devices which are called choppers. We might further restrict the subject, and confine our attention to devices handling rather limited power levels.

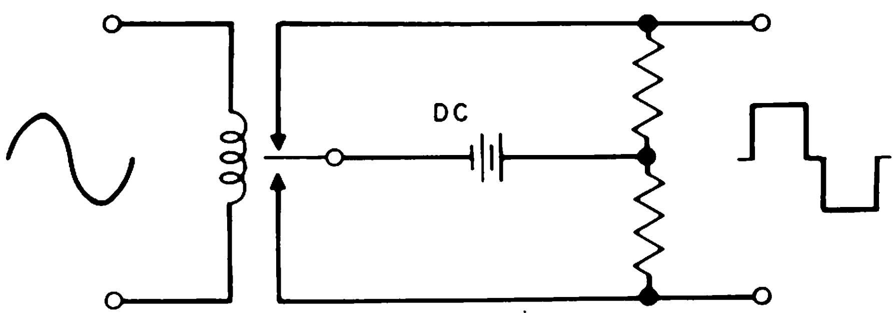


FIGURE 2. *Contact action measurement circuit.*

For evaluation purposes, the chopper will be considered to be driven by a perfect sine wave which is repetitive in nature. Because, by design, the contacts maintain a constant time relationship to the driving waveform, it is possible to completely describe the operation of a chopper by investigating one cycle. For convenience, we will speak of portions of this cycle in electrical degrees, 360 degrees constituting one complete cycle of operation. The output signal, for purposes of measurement, is defined as the voltage developed across resistors connected to the contacts and supplied with DC as in Figure 2.

TIME, as expressed in electrical degrees of the exciting frequency, is used as the basis of most expressions. Consider one complete cycle of a perfect sine wave used to drive the chopper, and reference all contact action to that cycle.

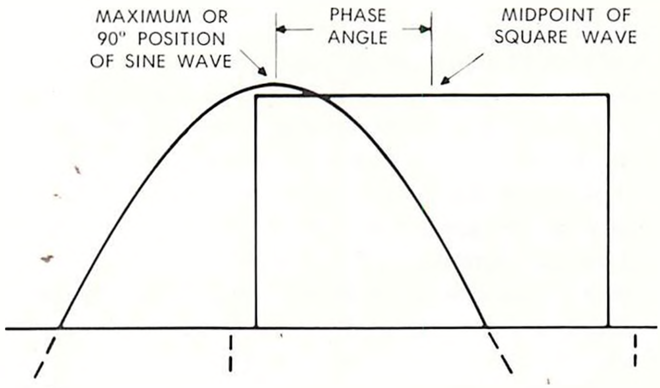


FIGURE 3. *Phase angle is measured to midpoint of square wave.*

PHASE ANGLE, the inherent phase lag, is defined as the angle existing between the peak of the driving sine voltage and the midpoint between contact make and contact break, expressed in degrees of the driving wave, as in Figure 3. That is, phase angle is measured from the 90° (or 270°) point of the driving sine wave to the midpoint of the on-time or period of closure. In a type 300 Airpax chopper at 400 cycles, this lag is roughly half electrical, (the coil L/R relationship), and about half electro-mechanical.

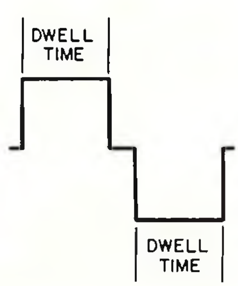


FIGURE 4. *Dwell time is period of contact closure.*

The drawing, Figure 3, illustrates the nominal phase relationship in a type 300 chopper at 400 cycles; for the sake of simplicity, the opposite half cycle is not shown.

DWELL TIME, also called on-time, closed-time, etc., is the number of degrees each contact is closed, expressed in relation to a driving sine wave, and is illustrated in Figure 4. Obviously, it can be and sometimes is expressed in milliseconds. This is inconvenient for measurement unless the chopper is used only at one frequency. At 400 cycles, a type 300 chopper has a nominal dwell time of 1.05 milliseconds; i.e., 147/360° x 2.5 ms.

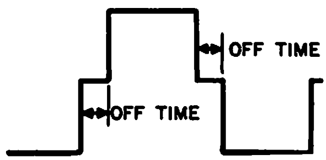


FIGURE 5. *OFF TIME of a BBM chopper is period when both contacts are open.*

BALANCE OR SYMMETRY. This is the difference between positive and negative dwell times; thus, if one half were 145° and the other 140°, the balance would be 5°. It is usually specified as a maximum (actually a maximum unbalance).

OFF TIME, transit time, or dead time. The period, in degrees of the driving sine wave, during which neither contact of a BBM (break-before-make) chopper is closed. See Figure 5. It is about 0.23 ms for a type 300 chopper at 400 cycles, and would usually be expressed as 33°; like dwell time, it occurs twice each cycle.

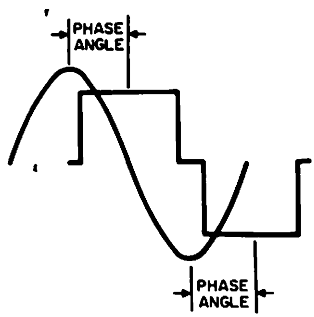


FIGURE 6. *A phase difference may exist between opposite wave halves.*

PHASE BALANCE refers to the possibility that the dwell times may not be perfectly symmetrical, and this might occur if the balance is in serious error. Figure 6, will clarify this. Usually this does not become a problem and choppers have phase balance within a degree or two even if the dwell time is seriously unbalanced.

RELATIVE PHASE, or polarity. If one of many chopper polarities, such as coil leads, is reversed, the chopper may appear 180° reversed. This would prevent many servo circuits from functioning. This is illustrated in Figure 7. Polarity is easiest to test and pretty definite if it is specified in terms of DC on the coil. A chopper is of course, a polarized relay, thus DC in one direction through the coil will always close the same pair of contacts (if they happen to be open). When the sinusoidal drive voltage is interrupted, an Airpax chopper armature comes to rest with the moving contact against either one of the fixed contacts.

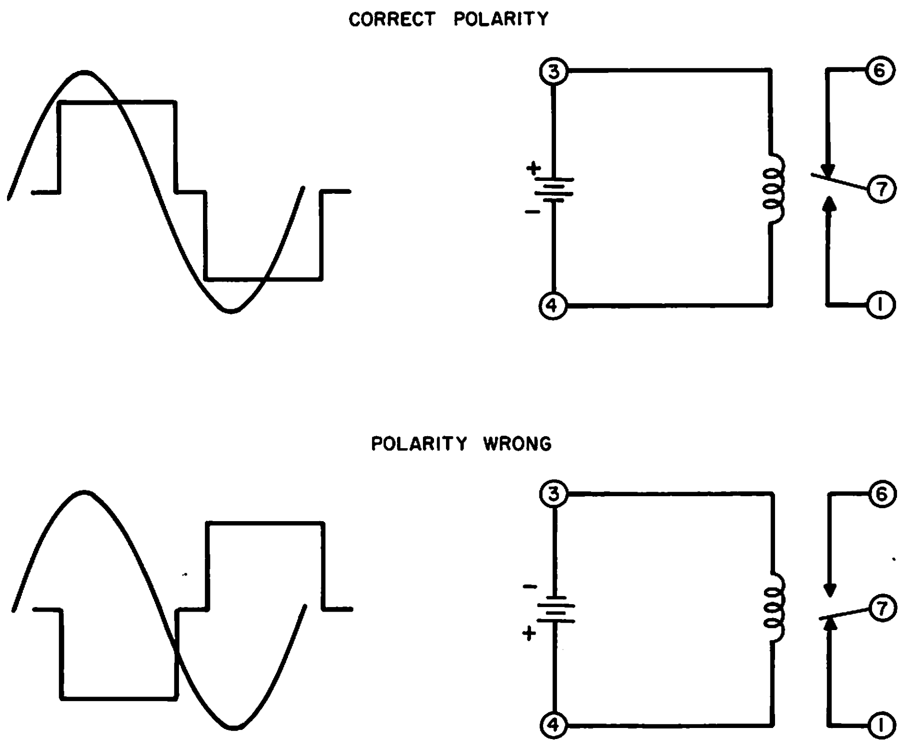


FIGURE 7. *Relative phase (polarity) of type 300 chopper.*

CLOSURE ANGLE is the angle between the sine wave and the beginning of dwell time, as measured on the base line. See Figure 8. This angle is not often used but is of importance in some circuits.

BREAK ANGLE. This follows from the definition of closure angle. It should be noted, if used, that this angle is governed by the dwell time.

COMMON TIME. In a make-before-break (MBB) chopper connected as shown in Figure 9, the output square wave pattern will be identical to the BBM type. In this figure, when all contacts are mutually closed, the algebraic sum of the voltages across the resistors is zero, identical with the condition when all contacts are open. Common time measurement provides a more accurate and easily measured control of balance. It is the converse of off time, and dwell time is actually the square wave length plus the common time.

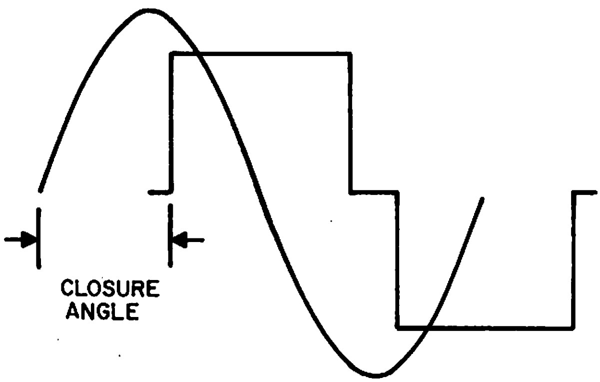


FIGURE 8. *Contact closing time is called closure angle.*

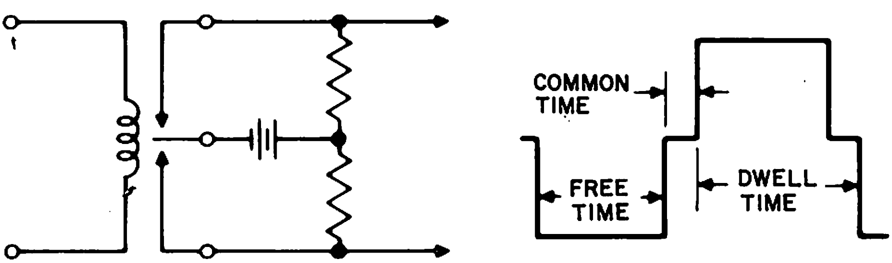


FIGURE 9. *In a make-before-break (MBB) chopper the sum of the common time plus free time is dwell time.*

FREE TIME is a term used to describe the square wave length of an MBB chopper connected as in Figure 9. Free time is that portion of dwell time when one contact is open; i.e., the operating contact is free of the opposite contact.

CHATTER, or contact bounce, Figure 10, usually appears quite close to the start or finish of dwell time; in fact, if it appears near the middle, it would indicate something wrong with the chopper. Chatter is rarely seen in an Airpax chopper; when it does appear, it will be found to occupy only a few degrees and is generally harmless.

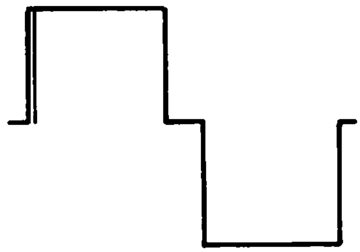


FIGURE 10. *Chatter, exaggerated in the above figure, is seldom observed in Airpax choppers.*

CONTACT DERANGEMENT refers principally to the phase modulation caused by mechanical vibration. It is defined, per Figure 11, as the aggregate of chatter, phase modulation and unbalance caused by vibration and is measured in degrees. Any Airpax chopper withstands vibration without damage; several types are designed to perform normally under extremes of vibration.

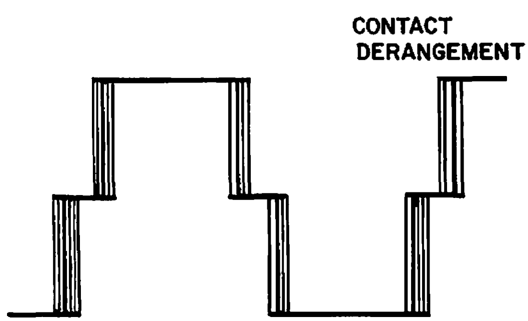


FIGURE 11. *Contact derangement is caused by vibration.*

NOISE is the undesired signal appearing between contacts and ground across the load resistor. It does not necessarily result in offset of a DC system, for one reason or another, as will be detailed in later sections. It is measured in rms, or peak, or peak-to-peak values. See Figure 12.

Noise is not specified properly until we have defined the circuit, the ground connections, specified the amplifier frequency range, defined the value as peak, average, etc., specified the meter used, which may be peak reading, average, or rms, and until we have called out the load resistance values. Finally, noise is still not specified until we can measure the noise originating in the chopper itself, and not in connecting leads or sockets or in nearby amplifiers.

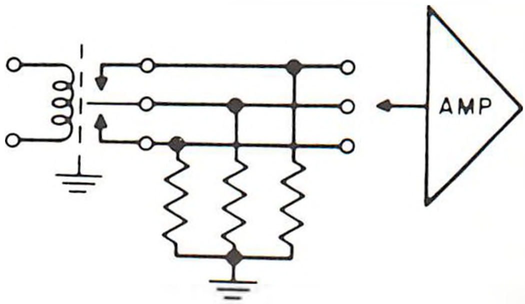


FIGURE 12. *Noise is the undesired signal from the contacts.*

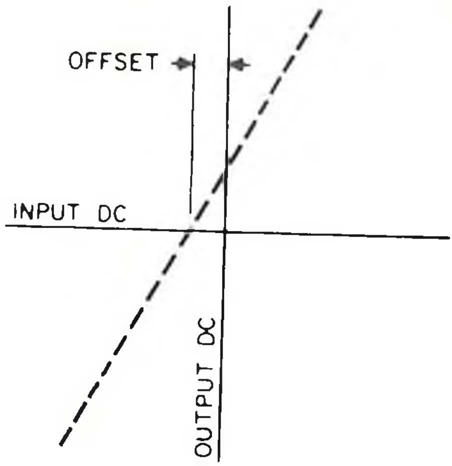


FIGURE 13. *Offset is measured as the DC input required to "zero" the amplifier output.*

OFFSET may be caused by the chopper, but is a function of the amplifier using the chopper. It is usually measured in terms of the amount of DC required at the amplifier input to return the amplifier output to zero. Offset is shown in the transfer function of Figure 13.

CONTACT RESISTANCE is caused by the appearance of powder between contacts, usually only when operated in a "dry", i.e., very low level circuit. When viewed on an oscilloscope, the pattern fluctuates erratically as the apparent contact resistance changes. The resistance is more likely to be a series of open circuits of very short time duration, similar to Figure 14.

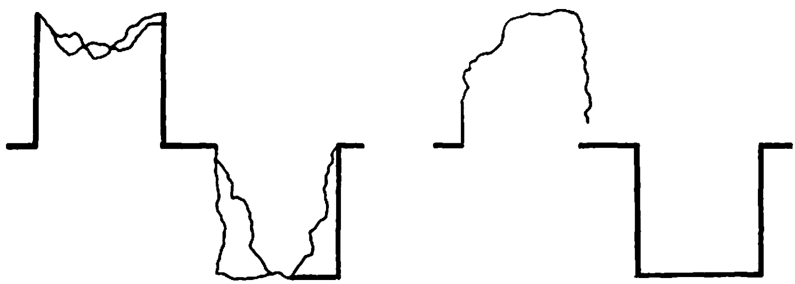


FIGURE 14. *Contact resistance shows as an erratic dwell pattern.*

LIFE. This one important factor often determines the use of a chopper. Choppers are sometimes regarded as unreliable, perhaps apt to quit working without warning. Such is absolutely not true, at least with Airpax choppers. Springs do break and coils do open up, but that kind of failure is so remote as to approach zero quantity. In the absence of other limits, a definition of the end of life is the time at which specification limits, as of phase or dwell time, are reached. As might be expected, the circuit used is important.

TEST VOLTAGES from contacts to ground are usually limited by an internal air gap to a maximum of 200 volts AC, but flashover causes no damage. The coil test voltages vary considerably, from 1500 volts for 100-volt coils to 200 volts on 6-volt coils. Insulation resistance is very high, being limited primarily by leakage across glass at the pins.

SECTION II

MEASUREMENT

Measurement of chopper parameters is usually done oscilloscopically. Unfortunately, most of the available phase meters depend on zero axis cross-over to measure phase. (The wave to be measured is converted to a square wave). The comparison of sine wave to step wave leaves these meters somewhat frustrated.

There are three possible means of measurement. We can display the pattern in rectilinear fashion and measure lengths with a ruler as in Figure 15, arrange the pattern in a circle, Figure 16, or mix a marker pip with the signal, Figure 17.

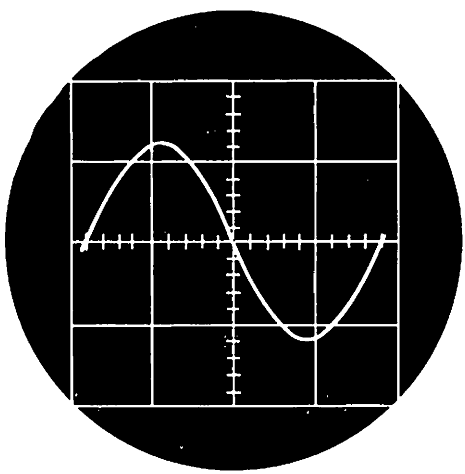


FIGURE 15. *Trace is several degrees in width.*

The rectilinear display of Figure 15, is the least desirable. The accuracy of the method obviously depends upon the linearity of the sweep circuit employed as well as the measurement of lengths.

Assuming a 5-inch scope tube is used with a trace 4 inches long, then 1 degree equals 4″/360° or .011 inches. If an accuracy in the order of 1 degree is to be obtained, it is necessary to make a linear measurement on the face of the scope tube accurate to .01 inch.

This is rather impractical if the curvature of the tube face and parallax error are to be considered. Also most sweep circuits employed in oscilloscopes are not completely linear. Therefore, if an oscilloscope is used, it is essential first, that the representation of 1 or 2 degrees be of a sufficient dimension to allow accurate measurement, and second, that the sweep be linear.

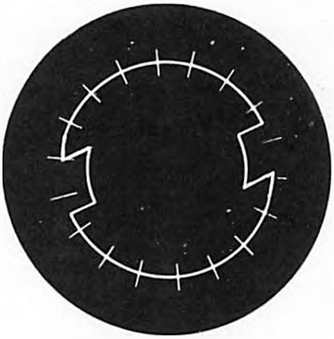


FIGURE 16. *Polar presentation provides a longer scale.*

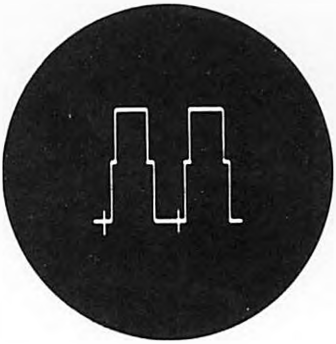


FIGURE 17. *Marker pips permit measurement of chopper parameters.*

Polar presentations provide a reading length π times longer, and also permit a 180° phase reversal to minimize error due to distortion and imperfect deflection accuracy. If a circle of 5-inch diameter is used, a trace length of 15.7 inches represents 360 degrees so that 1 degree equals 15.7/360° or .044 inch. This represents a minimum distance which can be practically read on a scope tube face. The linearity requirement can be readily fulfilled by using low distortion amplifiers, and obtaining an accurate 90-degree displacement.

No difficulty will be experienced in obtaining two voltages 90 degrees out of phase at a fixed frequency, and special circuits will yield the desired results over the expected frequency range of the chopper under test. The major errors occur in the CRT. The deflection plates are seldom exactly at 90 degrees to each other, and the circle becomes elliptical due to deflection distortion. Errors due to distortion can be practically eliminated by selection of the cathode ray tubes used in the measuring equipment.

After the circle has been established, it is simply necessary to drive the chopper from one of the two signals applied to the CRT and to interrupt the circle during the off time of the chopper contacts. The resulting pattern is shown in Figure 16. The foregoing method of dwell time measurement is the result of an extensive study to determine the most feasible system consistent with production practice.

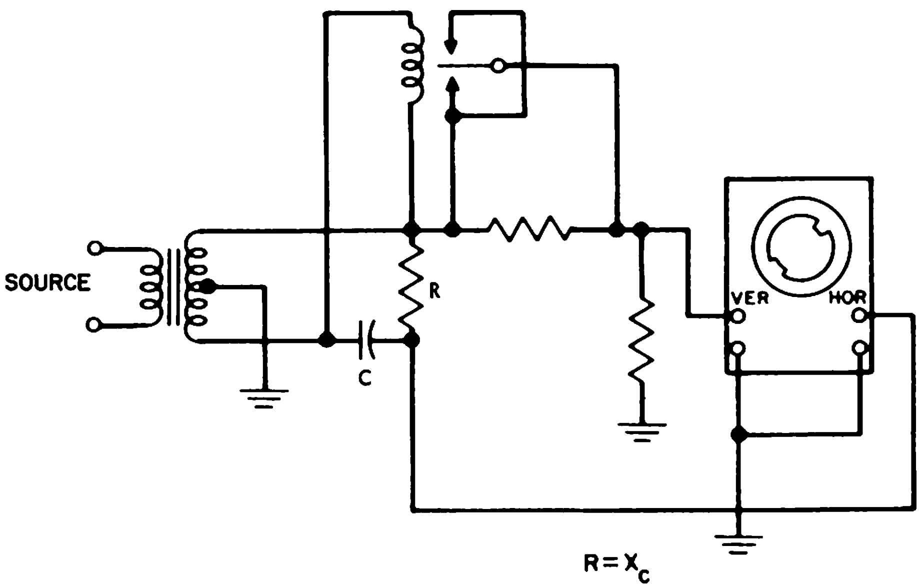


FIGURE 18. *A simple circuit permits polar display.*

A relatively simple circuit for a fixed frequency is shown in Figure 18. To measure dwell time, it only remains to place an azimuth dial over the pattern and to read this time directly in degrees. Some error will, of course remain, and can be further minimized. Referring again to Figure 16, if the two dwell periods are interchanged in time, errors in measurement can be directly observed. This can be done by reversing the polarity of the drive coil. Errors of measurement being reasonably the same on the same side of the scope, the measuring accuracy is improved.

Figure 19, depicts a rather elaborate piece of chopper test equipment, in which special effort provides low distortion amplifiers, push-pull deflection and other methods of obtaining a high order of accuracy. Signals having a 90° relationship are obtained with half-lattice networks as in single sideband transmission, and operation over a wide frequency range is therefore practical and easy.

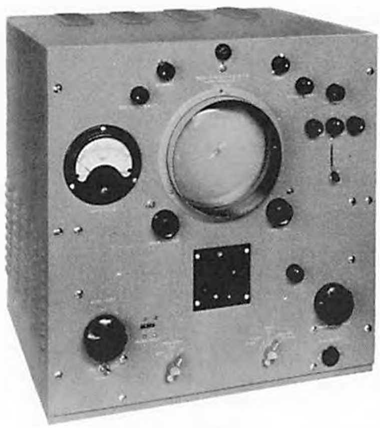


FIGURE 19. *Airpax production test equipment uses polar presentation.*

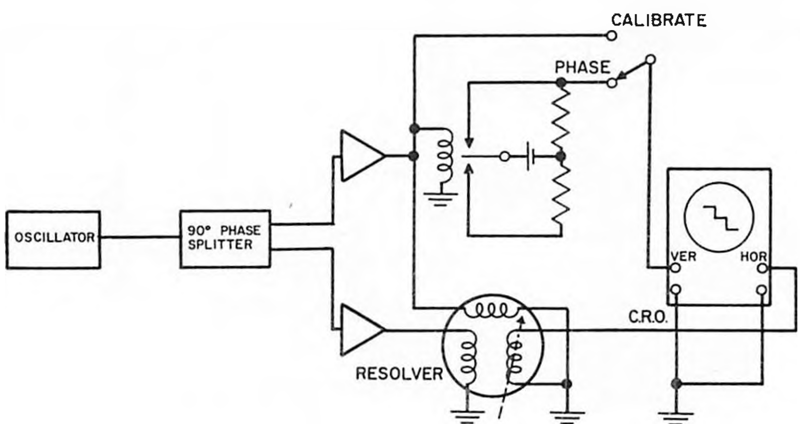


FIGURE 20. *A resolver simplifies measurement of phase angle.*

Phase angle is most easily read by the use of a resolver driven from a calibrated dial. In Figure 20, the output of a two-phase resolver is compared first against the chopper drive and then to the signal from the contacts, the difference angle being read on a dial. This point of closure of the pattern appears on the right in Figures 21 and 22.

The angular difference of the resolver shaft is the phase angle of the chopper. The accuracy of this method depends upon the accuracy of the resolver, low distortion waveforms, 90 degree displacement between the two voltages, and low phase shift in the scope amplifiers. All of these factors are controllable and hence, the accuracy of this method can be very good.

OPEN PATTERN CLOSED PATTERN

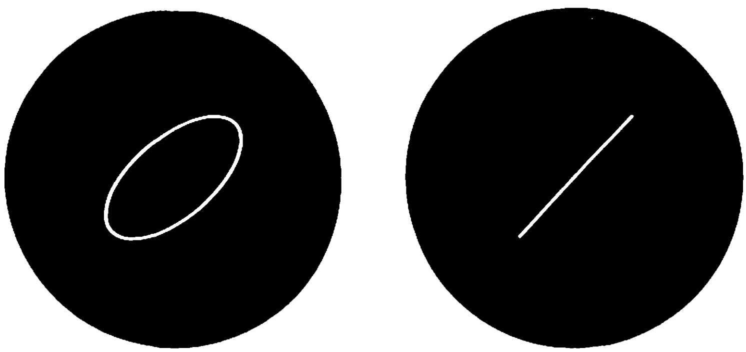


FIGURE 21. *Resolver position giving straight line is termed "zero".*

OPEN PATTERN CLOSED PATTERN

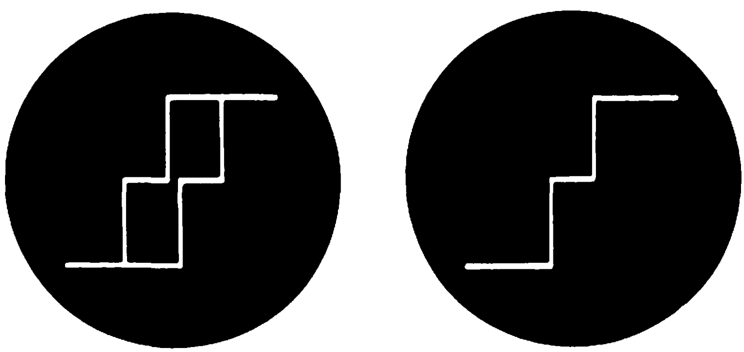


FIGURE 22. *Difference angle to close the step is chopper phase.*

Choppers from Airpax receive several complete tests before shipment, and all are given 50 hours of operation. (We thereby raise our quality level, eliminating any early failures). Results indicate this to be successful and our choppers exhibit a very high order of reliability and uniformity.

The following pictures show the adjustment and test equipment used at the Airpax Cambridge Division plant. Each chopper is put through exhaustive tests before final acceptance.

FIGURE 23. *A view of the "set-up" room where adjustments of phase and dwell time are made on choppers from production.*

FIGURE 24. *A close up view of two of the set up positions. Oscilloscopic patterns permit precise adjustments.*

FIGURE 25. *One of many final test positions. Tests include insulation resistance, noise and hi-pot on finished choppers.*

FIGURE 23

FIGURE 24

FIGURE 25

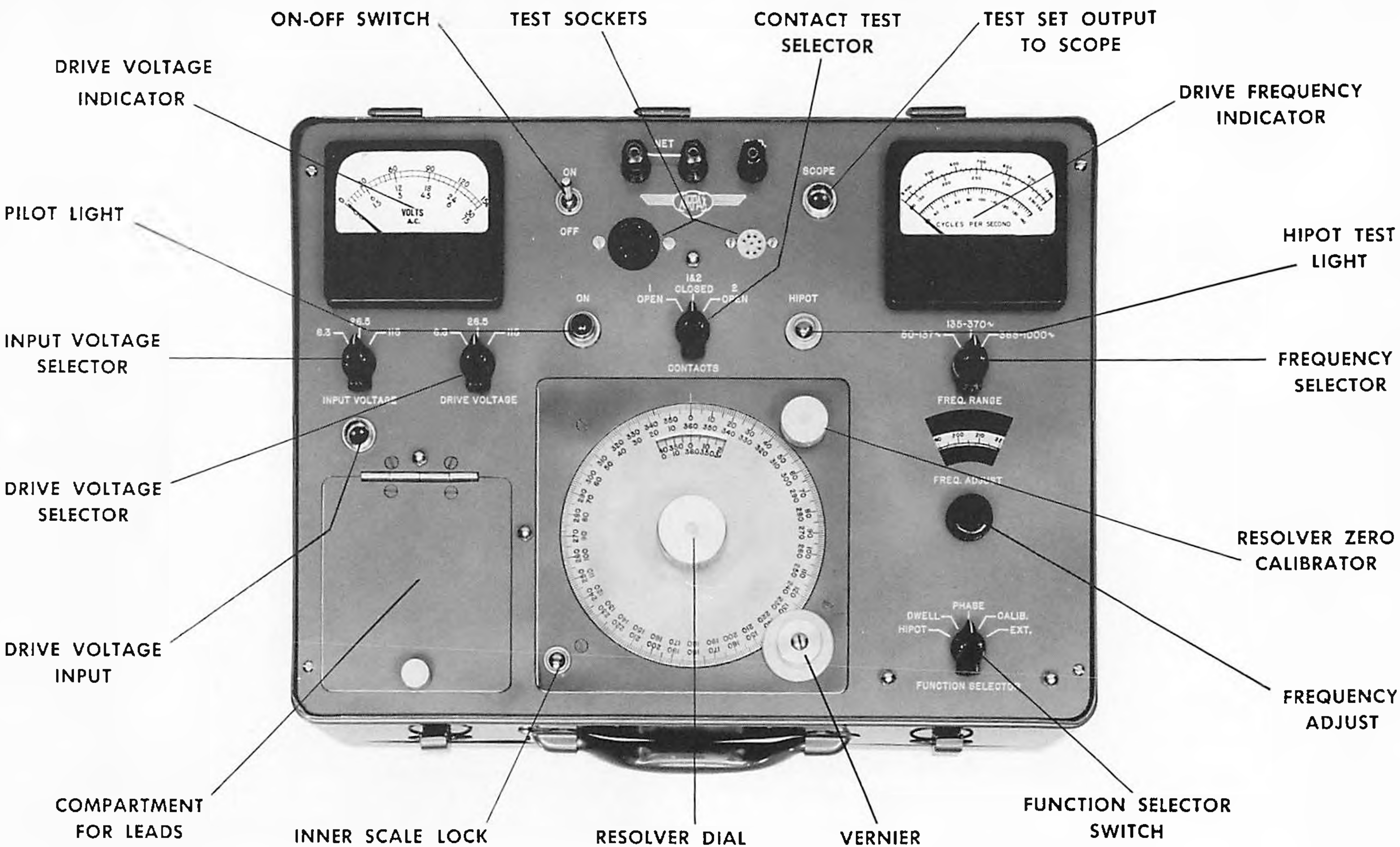


FIGURE 26. *Portable test set for chopper measurement.*

SECTION III

MEASUREMENT EQUIPMENT

The principle of having available a continuously variable phase permits the third type of presentation mentioned earlier, in which a marker pip is used to measure dwell time. A portable test set, Airpax TE-25, is shown in Figure 26, which gives a pattern previously shown in Figure 17. An external source of signal is needed to drive the chopper, and an oscilloscope to view the output. The equipment has a self-contained time base which connects to the horizontal amplifier. It is usable on any chopper to measure phase, dwell time, chatter, balance, and has a self-contained hi-pot test. The drive may be any voltage up to 150, or any frequency between 50 and 1000 cycles. The measurement accuracy is about 1 degree.

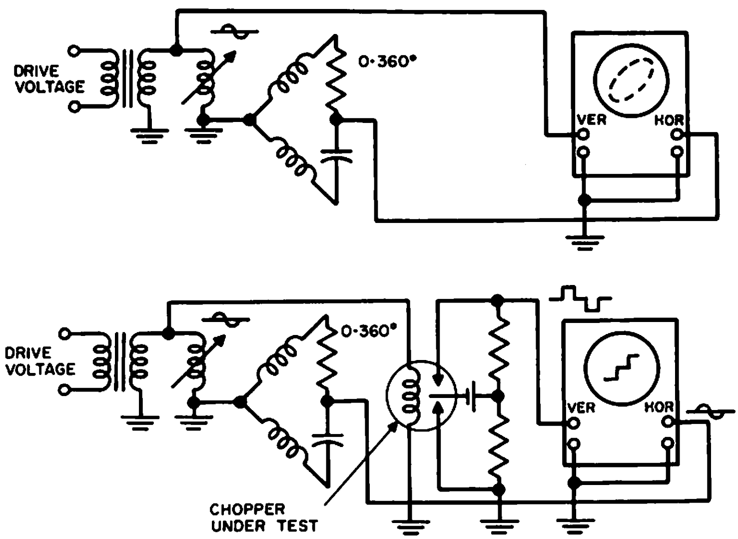


FIGURE 27. *Simplified phase measurement circuit.*

Figure 27, describes the basic circuits used for phase measurement. The two-phase resolver is supplied with a capacitive phase splitter, which is adjusted to 90° for each nominal frequency setting, a simple matter of setting a calibrated dial. The resolver provides highly accurate phase angles which are used for all measurements. To provide a simple method of zero adjustment, the housing of the resolver is rotated to zero, while shaft rotation supplies the measurement.

The measurement of chopper dwell time uses a very similar circuit, greatly simplifying the equipment and improving the accuracy thereby. In Figure 28, note that a marker pip generator is driven from the resolver output. After amplification, the resolver output is clipped and limited, integrated, and used to trigger a one pulse multivibrator. This pulse is again differentiated and the positive going pulse deleted. The spike finally obtained is mixed with the signal from the chopper contacts (or any signal). It can now be moved at will along the wave being examined.

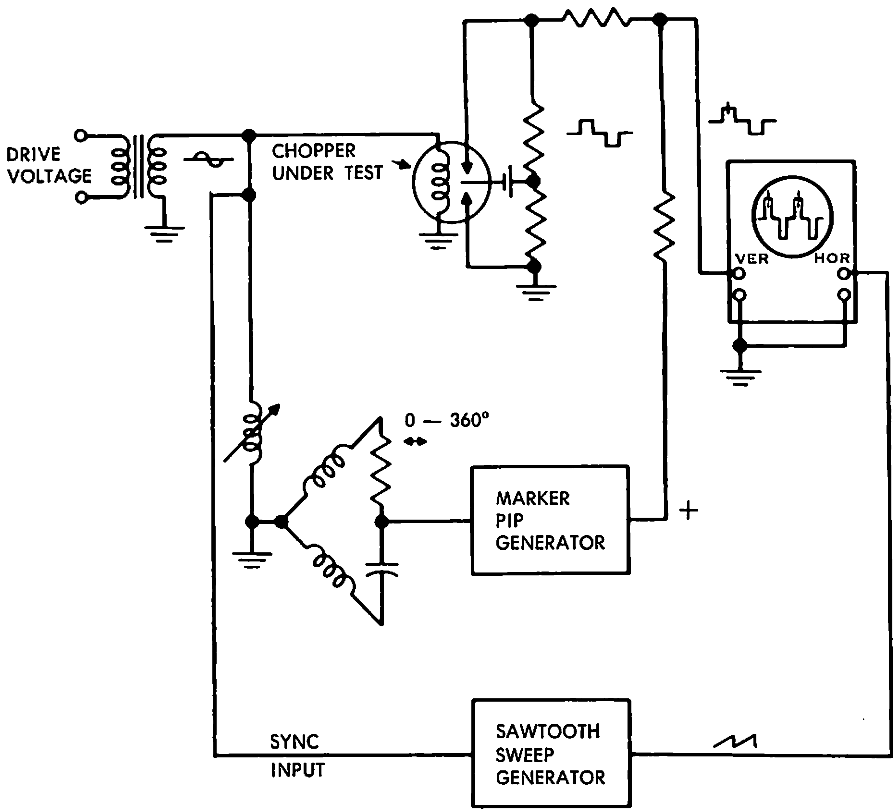


FIGURE 28. *Dwell time measurement circuit.*

It is, of course, quite practical to adjust the pip starting position in the same manner as with a phase measurement; however, it is inconvenient and time consuming to make this exacting adjustment. Avoiding this, the main dial carries a fiduciary or auxiliary dial which is locked and unlocked by a panel switch, thus "zero" or pip start can be at any dial position.

To provide a way of locating low insulation resistance or shorts, a 250 volt hi-pot test is self-contained, as in Figure 29. A neon bulb indicates a breakdown.

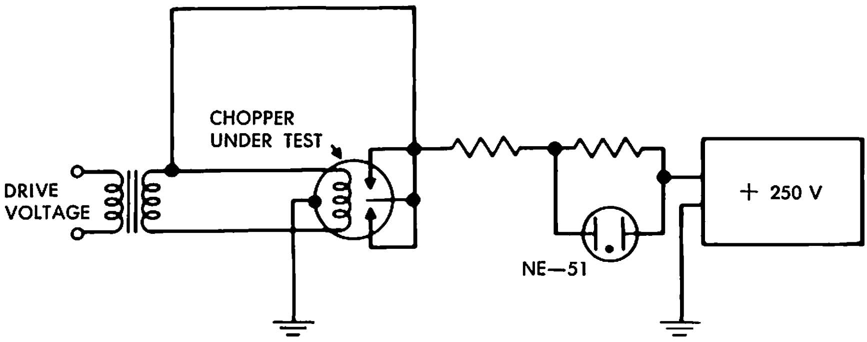


FIGURE 29. *Hi-pot test circuit.*

Model TE-25 is particularly useful for establishing phase angles between odd wave forms lying within its range, and Figure 30, shows the marker obtainable. Terminals are also provided to connect phase shifting networks ahead of a chopper, when it becomes necessary to adjust circuit or chopper phase.

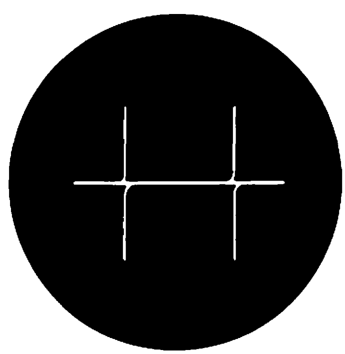


FIGURE 30. *TE-25 test set provides marker pips for various measurements.*

The method of adjusting chopper phase angle, using a resistor or resistor-condenser combination in series with the drive voltage, was described in PART I, page 7. It was shown that a chopper with a nominal phase angle of 65° can be conditioned to produce other phase angles.

Life testing is undoubtedly one of the most important chopper tests. Probably the life test circuit should simulate the circuit and conditions expected. Not knowing the application, we look for the worst. The worst seems to be dry circuit conditions and room ambients (apparently "hot" operation helps Airpax choppers, certainly it does no harm). It is important that the measuring circuits do not "clean up" contacts, which will happen if the current reaches as high as one milliampere. Operation with zero current seems to help powder formation, causing contact resistance, as mentioned earlier. Contact resistance due to powder formation produces irregular output pulses similar to a rapid series of open circuits. Irregularity is due to shifting of the powder particles as the reed vibrates. An oscilloscopic pattern of contact resistance was shown on page 8.

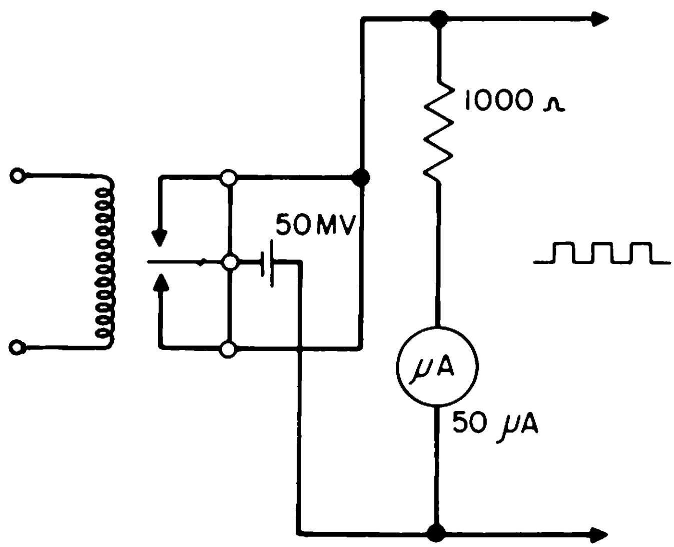


FIGURE 31. *Life test measurement circuit avoids cleaning up contacts.*

It is under these worst conditions that we find life figures greatly beyond 5000 hours, in fact, we have observed 25000 hours on Airpax choppers, operating at 400 cycles.

Figure 31, describes a test circuit measuring average dwell time. As any deterioration will be reflected by the contacts as a change, usually a reduction, of dwell time, this is a most sensitive measurement. Results can be easily read on a DC meter movement, recording or indicating, as it will integrate the pulse arrival to provide an average dwell time. It is also easily calibrated by shorting contacts (open the chopper drive lead).

Measurement is made at low level, 50 millivolts, 50 microamperes, to avoid a possible "cleanup" of the contacts. The 50 microampere DC meter permits visual observation if a recorder is not available. If the meter has a scale of 100, the reading will be about 80, the meter being set to full scale by shorting the contacts and adjusting the series resistor. (Shown as 1000 ohms). Contact resistance shows as erratic and reducing meter readings.

FIGURE 32. *Airpax life test equipment automatically examines choppers every few hours.*

Readings should be taken frequently when life testing choppers, and we firmly recommend not more than 24-hour periods. It is possible for contact resistance to appear, then disappear. Later tests may show an apparently good chopper but there may have been an open contact condition in the meantime. Figure 32, is an automatic life test equipment handling about 70 choppers. Once every few hours the dwell time is measured and recorded on strip charts. Complete records are kept and life test information is available for many years of testing. Once every hundred hours an operator examines all parameters of each chopper, such as phase angle, noise, dwell time, hi-pot, etc., which is kept with the recorded data.

Results of such records are illustrated in Figure 33, which shows the average dwell time of a normal chopper (top), and of a chopper which has developed excessive contact resistance.

Contacts which function best in dry circuits are usually not the best for relatively heavy currents. The effect of several milliamperes of current on a standard Airpax chopper is to slowly wear away the contact surface, reducing the dwell time noticeably. A model 300 chopper carrying 2 ma DC at 100 volts, for example, will wear enough in 1000 hours to get close to the lower specification limit of dwell time, and upper limit of phase angle. (If you have such an application, consult our Engineering Department. We can furnish choppers handling substantial power).

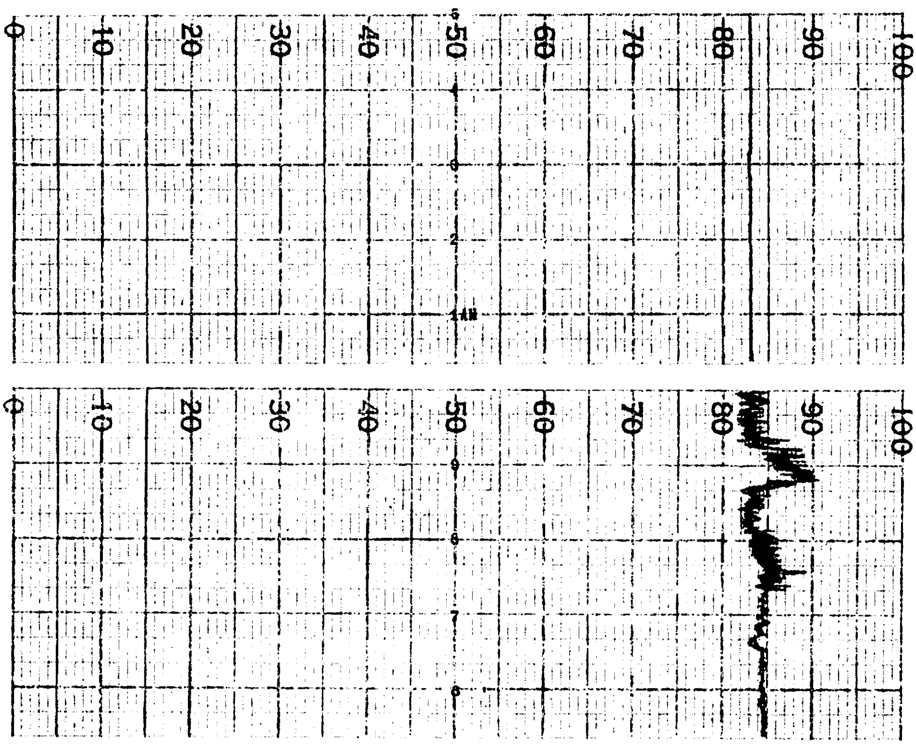


FIGURE 33. *Strip chart shows normal and erratic contact behavior.*

Noise has been referred to as contact resistance, previously shown in Figure 14, or as amplifier offset. By our definition, as described earlier, it is the stray or unwanted signal appearing at the contacts, when there is zero signal input. It must also be noted that the measured amount of noise may bear only a general relationship to resultant offset of an amplifier, due to the noise phase relative to signal, and other factors such as bandwidth.

A fundamental limitation[1] is set by the noise of thermal agitation, or Johnson noise, which is defined by the relationship:

$$E_{eff} = \sqrt{4KTRF}$$

where $K = 1.37 \times 10^{-23}$, T is degrees Kelvin, R the resistance involved, and F, the bandwidth.

If the resistor has a value of 1 megohm and a wide band amplifier of 100 KC is used, at room temperature (300°K), the noise level will be 45 microvolts effective value, or about 360 microvolts peak-to-peak. This applies to wire-wound resistance values, carbon resistor noise is likely to be considerably higher.

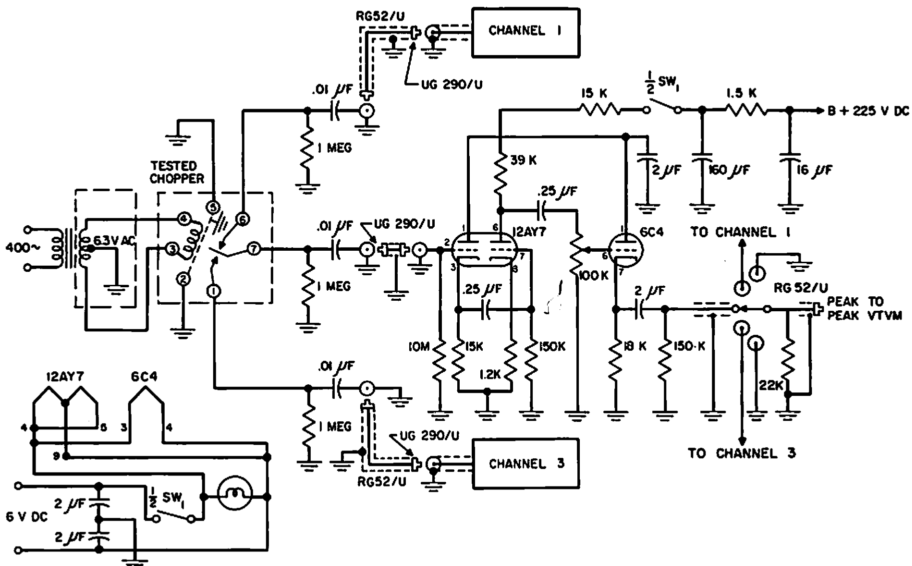


FIGURE 34. *Airpax production noise test circuit.*

Noise levels above 100 microvolts can be measured with the circuit of Figure 34, which is production test equipment used at Airpax.

The impedance level from chopper contact to ground, where the measurement is made, is in the order of 1 megohm. The amplifier output is low impedance to reduce hum pickup and to minimize the effect of switching transients on the VTVM. The circuit of the equipment is conventional, but special precautions are observed to produce a low initial noise level. Drive voltage leads to the chopper socket are well shielded. The driving source is balanced to ground. Precautions are taken to assure that the vibration of leads produces negligible electrical voltages. The amplifier has a flat frequency response from about 20 CPS to over 100 KC. Hence, any voltage appearing within this range will have its peak-to-peak values indicated on the meter.

[1] DC Amplifier Stabilized for Zero and Gain , Williams, Tarpley and Clark, AIEE Transactions, Vol. 67, 1948.

This is a rigid test to be imposed on a chopper, and in general provides a pessimistic measure of the chopper noise obtained in most circuit applications. Practical circuits may not have such a wide frequency response and may only be conscious of voltages over a restricted range. Chopper noise level also varies with the circuit impedance used at the contacts. The illustrated circuit is applicable only to the lower voltage drives, as 6 volts, and is shown connected for Airpax type 300.

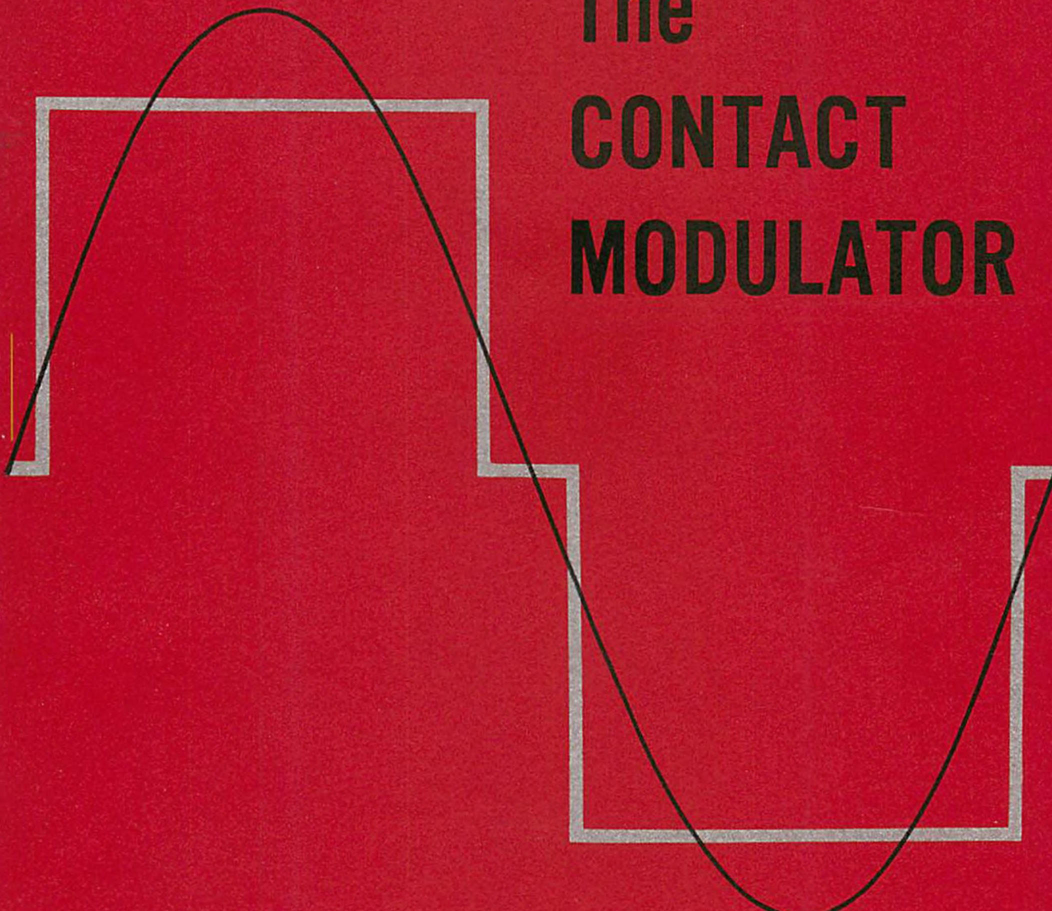
Below about 100 microvolts of noise we are usually interested in relatively low impedance value. If measurement below 10 microvolts is required, an excellent test amplifier is the Volker and Schafer Model VS-64A, a "hushed" transistor amplifier having a minimum noise level a little below one microvolt, for low impedances and a rather restricted bandwidth.

Following sections, in particular those discussing Application and Theory, further elaborate on noise problems and circuits to obtain optimum performance.

AIRPAX

Cowles.

The CONTACT MODULATOR



PART 2 Modulation Methods and Applications

THE SECOND OF A SERIES

AIRPAX

THE
CONTACT MODULATOR

PART II
MODULATION METHODS and APPLICATIONS

SECTIONS I, II, III, IV, V
WRITTEN BY

I. C. HUTCHEON
Bedfordshire, England

•

SECTIONS VI, VII, VIII
WRITTEN BY

H. A. COOK, President

and

THE ENGINEERING STAFF

of

AIRPAX ELECTRONICS
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TABLE OF CONTENTS

SECTION I

SOME BASIC PRINCIPLES

Mechanical choppers were developed originally for converting very low level DC signals into AC, so that drift-free DC amplification could be obtained using simple AC-coupled amplifiers, and this is still their prime function. Many other applications have since been found such as demodulation, time-sharing and DC isolation, to quote three examples, but modulation is still the most important.

Any device, which in the broadest sense will multiply, can be used as a modulator, and several have been outlined in Part 1 of this series. The mechanical switch, however, is especially efficient since it can multiply a signal by very nearly zero or very nearly unity, and this is because its open resistance is very high, its closed resistance very low, and its noise (in a well designed unit) extremely small. This Part deals with modulators based on the use of switches. Transistors also make quite good switches, and providing their limitations are borne in mind, many of the following remarks apply to circuits using them as well.

Switch-type modulators convert a direct input signal into a nominally rectangular AC carrier wave at the chopping frequency. This "square wave" can then be amplified by a relatively simple AC-coupled amplifier, and reconverted to DC at a higher level by a demodulator and filter. If the input changes polarity, the carrier reverses its phase and the demodulated output changes polarity as well. The response of such a system to alternating or step inputs depends on the design of the modulator and the demodulator. If both are full-wave and accurately synchronized, for example, and the off time is short, the input signal frequency may approach that of the carrier. Signals above carrier frequency are also transmitted faithfully during the on times. In many practical circuits, however, the use of half wave circuits and the possibility of some drift in contact adjustment make it desirable to smooth the output, and this will limit the response to some fraction of the carrier frequency. Although some sort of demodulator is invariably necessary to extract the information from the carrier, it need not have an electrical output. For example, an AC servo motor is a demodulator whose output is angular velocity. One of its inputs is the AC carrier, and the other is an AC reference of the same frequency.

When the output of a chopper-type DC amplifier is zero, the input should also be zero, but in practice many factors such as pickup in the signal, thermal emf's in the wiring, amplifier hum etc., conspire to ensure that it is not. The input signal which gives zero output signal is termed the offset or zero offset of the amplifier. Its magnitude is

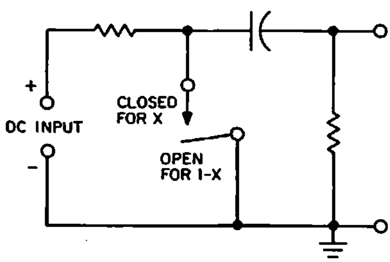


FIGURE 1.
On-off modulator.

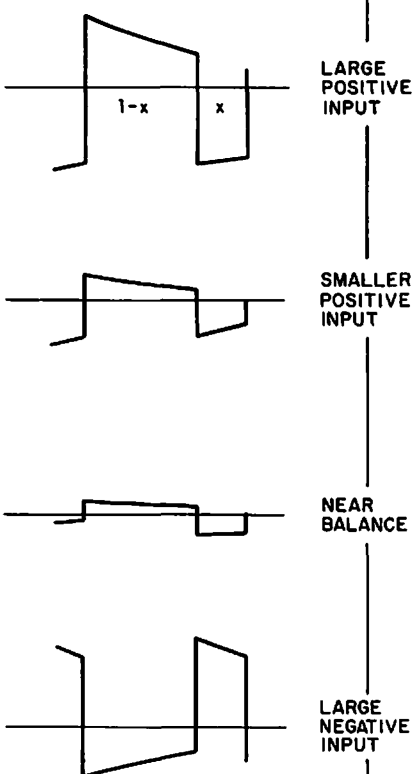


FIGURE 2. *Output waveforms of on-off modulator.*

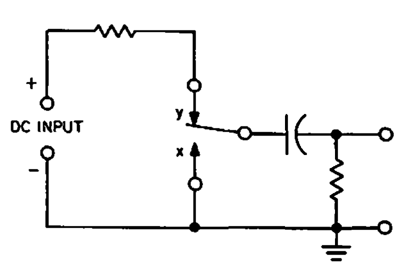


FIGURE 3. *SPDT modulator (BBM).*

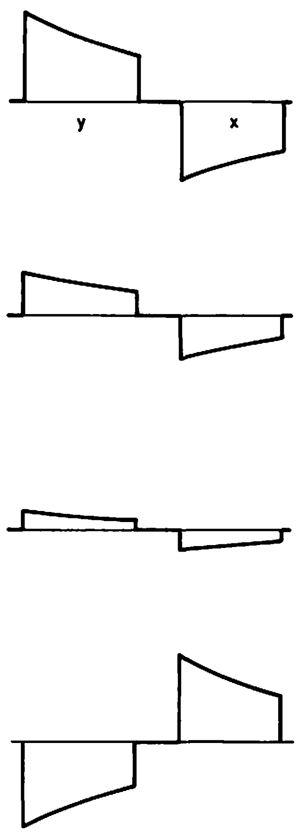


FIGURE 4. *Output waveforms of SPDT modulator.*

important since, in the last resort, it determines the accuracy of amplification. Some components of the offset are stable and respond to compensation; others, usually known as drift, comprise predictable components which can be compensated and random components which cannot. In a carefully designed amplifier, the offset usually depends on the noise generated in the chopper itself, and some commercially available types of mechanical choppers have voltage and current noise levels below one microvolt and 10^{-11} amperes respectively.

Although the offset can be made very low, the gain stability of the modulator—AC amplifier—demodulator system is seldom very good (except in a few special circuits), being dependent on waveform mark-space ratio and other factors. For this reason, negative DC feedback is almost invariably applied around the whole system to stabilize the gain. The modulator then handles a difference or error signal which is kept very small by the action of the feedback. The forward gain is made so large that full output is obtained for an error signal which may be no more than e.g. 1% or 0.1% of the input signal. As the gain is increased, the accuracy improves in proportion until the error signal has been reduced to about the same level as the modulator offset. Any further increase in gain gives no advantage and has the disadvantage that it renders the system less stable, and more prone to saturation by e.g. stray pickup signals.

The error signal may be derived first and then applied to the modulator as a single input, or alternatively the modulator may itself respond to the difference of two large signals applied independently. Very often one of the input signals is fed through a low-pass filter to remove unwanted pickup or noise. The feedback from a demodulator, however, is usually smoothed by a single RC stage to ensure stability of the feedback loop, and it is therefore applied without additional filtering which would provide a second lag.

A simple modulator circuit is shown in Figure 1, and provides a nominally rectangular waveform which may be symmetrical or not as desired. Typical waveforms are illustrated in Figure 2, which also shows how the output changes phase as the input changes polarity, and is zero when the input is zero. Figures 3 and 4, show a rather more efficient circuit with a somewhat different waveform.

Practical waveforms differ somewhat from the ideal, and there is usually some droop, as shown, due to the finite time constants involved. But it can be made small, and we shall ignore it. The rise of leading edges, and the fall of trailing edges can also be delayed if the load contains significant shunt capacity, or the amplifier response is poor, but again this is a secondary and usually avoidable effect and we shall ignore it too.

SECTION II

PRACTICAL MODULATOR CIRCUITS

So much for the basic principles. What is required from a practical modulator, and how is it best obtained? Here are some of the factors that may need to be taken into account.

DESIGN CONSIDERATIONS.

1. Need for isolation between DC input and AC output.
2. Need to compare two input signals.
3. Need to compare two inputs at different potential levels.
4. Degree of immunity required to pickup, both series and common mode.
5. Effects of amplifier noise, hum and low frequency drift.
6. Stability considerations when there is overall DC feedback.
7. Source resistance and amplifier input impedance.
8. Output waveform desired.
9. Type of chopper available.
10. Conversion efficiency.

It is interesting to consider some practical circuits with these factors in mind.

SIMPLE GROUNDED CIRCUITS. The circuit of Figure 1, is a useful one when the input is initially floating, can be grounded on one side to the amplifier, and does not contain too much pickup. It is economical in contacts, and allows the other half of an SPDT chopper to be used to demodulate the amplified signal, or to act as modulator in a second amplifier. One or other of these arrangements is common in the subsidiary chopper amplifiers used to stabilize direct-coupled amplifiers for analog computer use. The circuit is not particularly efficient, since input current is drawn to no useful purpose every time the switch closes, and for this reason an on-to-off ratio less than unity is sometimes used.

More efficient is the SPDT circuit of Figure 3, which eliminates the wasteful drain by using another contact. Efficiency is not usually a very important consideration however, unless the limits of noise are being approached. Of greater importance is the fact that both these circuits are very susceptible to pickup at the chopping frequency, this being at least partially passed on to the amplifier and so causing a DC offset. Thus, both are likely to be unsuitable where there are long input lines, such as occur in thermocouple temperature measurement, unless the chopping frequency differs from that of the supply mains.

COMPARING TWO INPUTS. There are in principle two ways of comparing two DC signals so that the modulator output depends on their difference.

The simplest and most accurate is to connect the two signals in series and apply the difference as a single input to circuits such as those in Figures

1 and 3. In this case one of the inputs is operated at a varying potential to ground, assuming the load to be grounded. However, a slight modification (Figures 5a and 5b), in the connection of the load resistor to the DC carrying loop of the circuit permits both inputs to operate with one side at ground potential. When the two inputs are equal, the coupling capacitor becomes charged to the common potential and there is no AC output. Either input can be filtered.

In the other method, shown in Figure 6, the two inputs must be of opposite polarity to ground, in contrast to the above, and are summed as

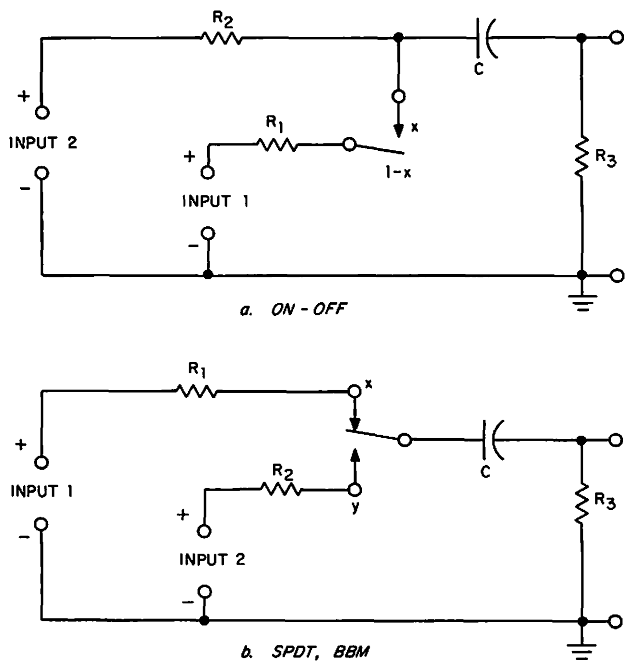


FIGURE 5. *Two grounded inputs can be compared in series.*

currents through two accurate resistors onto one point. Sources of low or known resistance must be used, and the accuracy of comparison depends on that of the resistors.

A CAUSE OF LOOP INSTABILITY. The circuits shown in Figure 5, may be a subtle cause of instability in cases where one of the inputs is in fact a feedback signal derived from the amplified and demodulated output.

Consider either circuit and assume that both inputs are suddenly raised by the same amount. Since the circuit time constants are several cycles long, the coupling capacitor takes a few cycles to charge up, and the AC output signal meanwhile takes one of the forms shown in Figure 7.

Both signals contain an AC component and a change in DC level, and their effect depends on the design of the demodulator. A balanced demodulator will respond mainly to the AC component, but a one-sided type may respond to the DC shift as well.

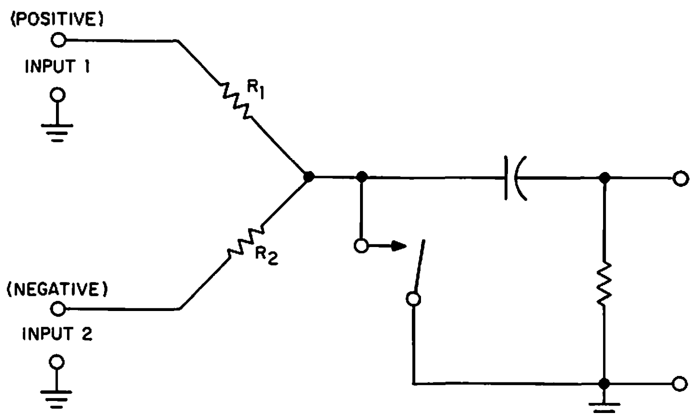


FIGURE 6. *Two inputs can be summed through accurate resistors.*

The essential requirement for stable operation is that the total effect should be to reduce the final DC output. Any tendency to increase it is equivalent to a lag in the feedback, and will tend to cause oscillation. Thus in the circuit of Figure 5a, if the demodulator recognizes only the AC

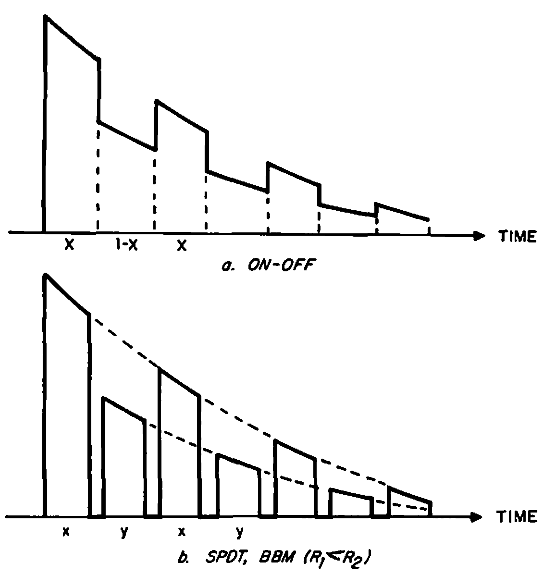


FIGURE 7. *Output due to equal step change in both inputs to the circuits of* Figure 5.

component, the feedback signal should be introduced at input No. 1, and R_1 should preferably be small compared with R_2. The AC signal then has the phase associated with too large a feedback signal, and so tends to reduce it. Injecting the feedback at input No. 2 in this circuit is bound to give the effect of a lag. In the SPDT circuit, the resistor associated with the feedback input should be the lower of the two.

The difficulty is avoided if the two inputs are compared first, either in series or by parallel summing through resistors, and then fed in as one input, provided that no filtering is employed.

An ideal arrangement, though a more expensive one, is to use a fully

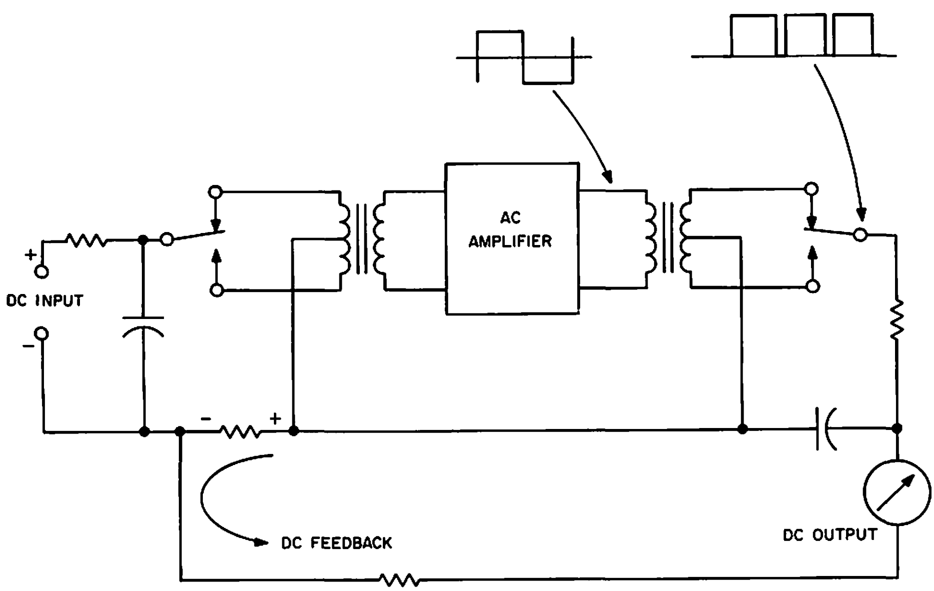


FIGURE 8. *Fully balanced system minimizes instability problems.*

balanced modulator such as the SPDT transformer-coupled circuit shown in Figure 8. The amplifier then handles pure AC, with no changes in DC level even during transient conditions. Use of a balanced demodulator is an added safeguard against unbalance developing in the modulator. It also eases the smoothing problem, and it is worth noting that, in the ideal case, with perfectly synchronized demodulation, zero transit time in the switches, and no droop in the waveform, no smoothing is required at all.

PICKUP. Unwanted AC signals reaching the input of a chopper-type DC amplifier may seriously affect its accuracy, either by causing a DC offset or by blocking the later stages of the AC amplifier. They are classified under two headings.

1. Series-mode AC pickup is the description given to any alternating voltage appearing in series with the DC input signal. It is frequently caused by the interaction of stray alternating magnetic fields with circuit loops formed by the input leads and the DC source, and its effect can be reduced by suitable filtering before the modulator. Filters inevitably degrade the response of the amplifier, however, and their design generally involves a compromise between the requirements for high AC attenuation and fast response. Where the pickup is at the chopping frequency, balanced modulators are to be preferred, since their output is at twice this frequency and no zero offset results. Saturation may be caused by pickup of any frequency, and the tendency for this to occur can be minimized by demodulating at the lowest possible signal level. This is achieved by providing as much as

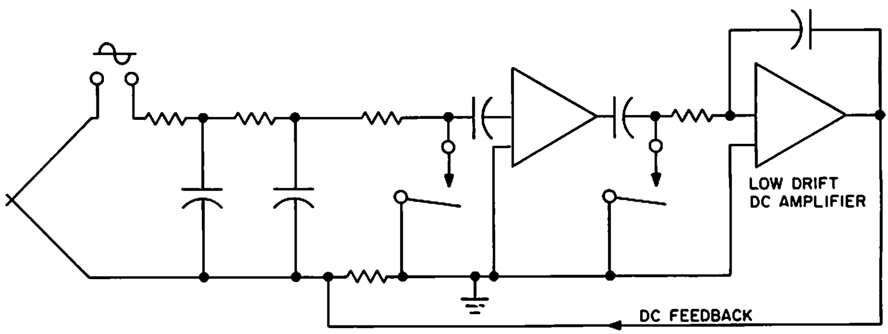


FIGURE 9. *Input filter and low level demodulation reduce effect of series AC pickup.*

possible of the required forward gain by direct-coupled amplification after the demodulator and smoothing circuit, so that the gain of the AC amplifier is as low as possible. The limit to this procedure is set by drift in the demodulator and the DC amplifier. An arrangement of this type is shown in Figure 9.

2. Common-mode signals are those which are applied between the input terminals and ground. They may be current signals of high source impedance arising, for example, from capacitive coupling between the input leads and the supply mains, or they may be voltage signals of low source impedance. They may be AC or DC. In the case of current signals which are not too great,

it may be possible to use a grounded amplifier having a simple capacity-coupled modulator as shown in Figure 9. If the current is very large, however, as may arise when a thermocouple is situated inside an electric furnace, or if the signal is essentially a voltage one, it becomes necessary to float either the whole amplifier or the input and feedback circuits at the common-mode potential. Both methods are used. The former makes it necessary to provide floating screens for at least the more sensitive parts of the amplifier. The latter requires the modulator to be transformer-coupled to the amplifier, and the feedback circuit to be isolated from the amplifier output.

One method of providing the necessary isolation at the output end of the system, employed in many self-balancing potentiometric recorders and shown in Figure 10, is to use a motor-

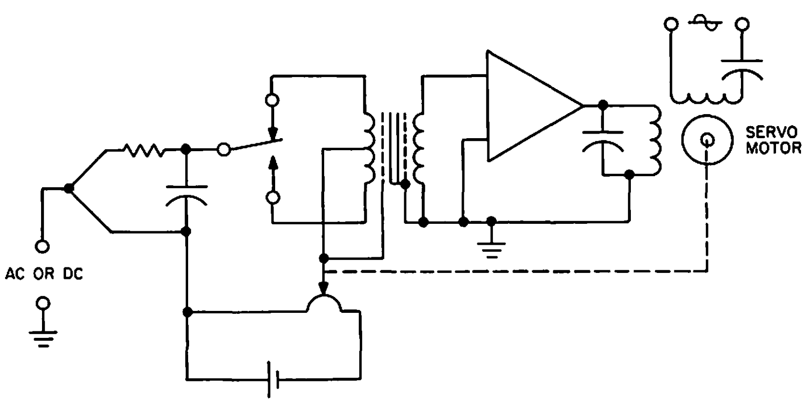


FIGURE 10. *Transformer coupling gives common-mode rejection.*

driven slidewire fed from a dry battery or other floating reference supply. Another method is to use a DC isolating device based on magnetic principles. Transformer coupling of the modulator has the obvious advantage of permitting good impedance matching between the input circuit and the amplifier, as well as providing a balanced circuit which gives useful rejection of series pickup.

INPUT TRANSFORMER DESIGN. Transformers suitable for very low level operation are rather special, and their design should not be undertaken lightly. A high permeability nickel-iron core is used, and every winding, including half windings, must be wound in two equal

parts on opposite legs so as to cancel out induced voltages. Even so, a high permeability external shield is almost always essential, and two or more shields with air spaces in between are not uncommon for more rigorous applications. One commercial design uses two 1/32 in. thick Mumetal shields and has a shielding efficiency of 60 db. To avoid loops outside the screen, the terminations are usually fly leads in twisted pairs, not tabs or terminals. The effect of common-mode AC pickup is directly proportional to the inter-winding capacity, and efficient screens must be incorporated. Transformer inductance should be measured with small signals, since the initial permeability of the core may be several times less than the large signal value. In extreme cases it may be necessary to demagnetize the entire assembly to prevent microphonic effect, i.e., signals induced by the flexure of the whole structure. The literature contains several useful references on transformer design for this type of application, and there are several models on the market.

DC ISOLATION. Sometimes it is necessary to compare two signals which are at different levels relative to ground, and almost the only solution is an isolating circuit using a floating capacitor switched between one part of the circuit and the other. Figure 11, shows an arrangement which finds at least one commercial application in the measurement of pH, the input signal coming from two electrodes which are one or two volts away from ground, and requiring balancing by a feedback signal which is at ground potential. Transformer coupling and a floating feedback are not practicable in this application, on account of the high resistances involved. For steady state analysis, the isolating circuit reduces to the simpler equivalent circuit shown in the same figure.

There is something to be said for operating this system at low frequencies. The transfer capacitor and its associated contacts inevitably have small capacities to ground and to the other contacts, and these are repeatedly charged to the potential of the input circuit and then discharged again, this happening once a cycle. The effect is to draw through the source resistances, currents which are proportional to frequency and which alter the potential across the filter capacitor, so causing an offset. This error is reduced proportionately by reducing the frequency, and the pH system mentioned previously operates at around 1 CPS, the chopper in this case being a relay with special contacts and very good insulation. Source resistances up to 1,000 megohms are permissible, and accuracy is better than 1% with an input range of 0 to 100 mv.

THE OUTPUT FILTER. Whether an amplifier contains valves or transistors, a direct potential exists between its input terminals, and some form of AC coupling is essential to prevent small direct currents from reaching the chopper. Any form of DC leakage to the chopper is likely to give rise to an AC output, and leakages as small as 10^{-11} ampere may

cause trouble in sensitive modulators. Hence the coupling capacitor must be of a suitably high quality. If the current sensitivity of the system is high, or the chopping frequency low so that a large actual capacity is needed, it may be impossible to achieve the desired insulation resistance in a single capacitor, and in this case a two-stage CR output filter will solve the problem. The voltage across the first capacitor, due to leakage through the second, is so small that leakage through the first one can be neglected.

Precisely the same trouble can be caused by slow variations in the potential between the amplifier terminals, even if the capacitor is perfect. Heater voltage changes in valve amplifiers, or temperature-induced

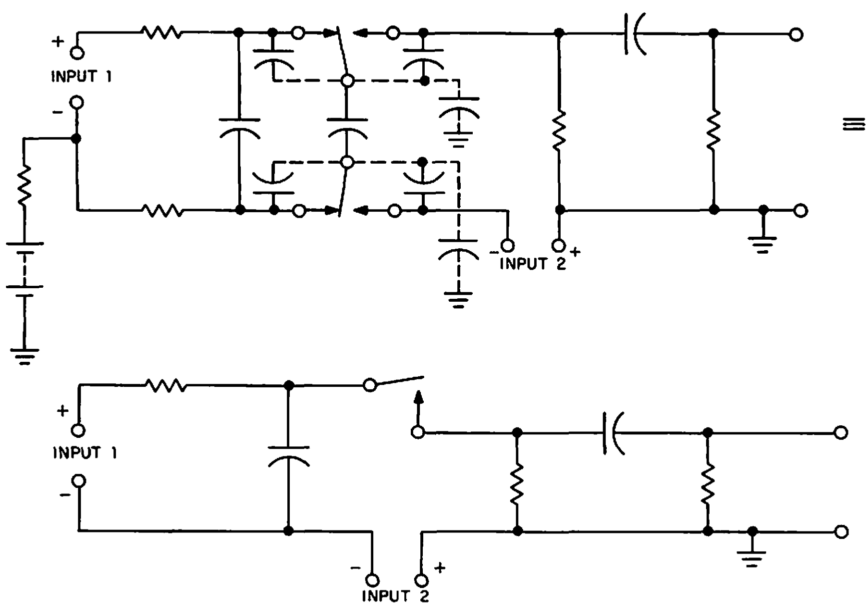


FIGURE 11. *DC isolating circuit is equivalent to grounded circuit in the steady state.*

changes in base to emitter voltage in transistor amplifiers may dispatch slowly varying currents to the chopper, causing in some cases disastrous temporary swings in the demodulated output. The solution is the same, namely to use a two or three stage high-pass filter which transmits the chopped signal forward to the AC amplifier but attenuates slow changes in the opposite direction. The arrangement is shown in Figure 12, and in the steady state is equivalent to a single resistance load whose value is equal to the filter resistances in parallel. Balanced input arrangements are considerably better than one-sided circuits, and symmetrical transformer-coupled arrangements are fairly immune.

Very occasionally, if the first amplifying stage is, for example, an electrometer valve having extremely low grid current, the coupling capacitor may be dispensed with altogether. Modulation is still useful in eliminating the slow DC drifts of such an amplifier.

BBM OR MBB? Single-pole double-throw circuits (and two-pole changeover circuits in the case of transformer coupling) can be either break before make or make before break. So far, BBM has been implied, but each system has its advantages.

Capacity-coupled BBM circuits are most efficient when a filter capacitor is connected across the input (Figure 13). The amplifier then sees a low

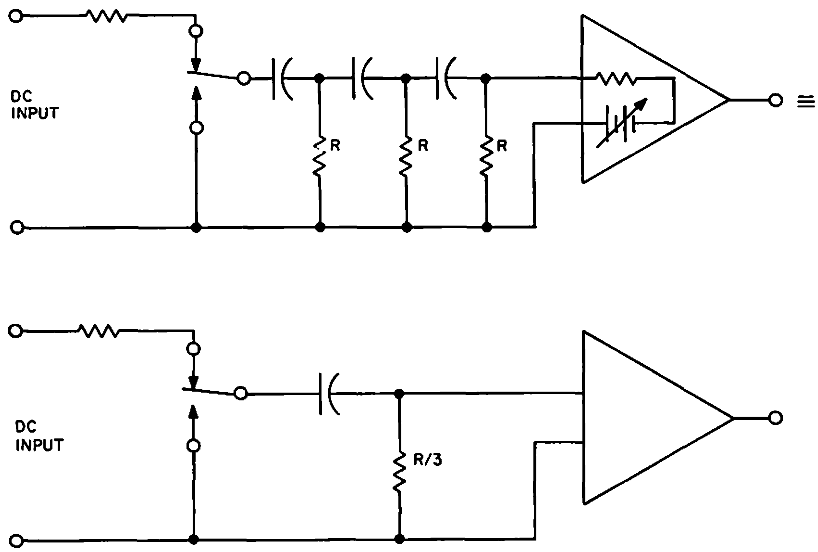


FIGURE 12. *High-pass output filter reduces effect of potential variations at amplifier input.*

AC impedance to ground at all times except during the transit time of the chopper, and this may permit spikes to appear in the output signal. The MBB setting avoids this trouble, provided that the chopper contacts remain perfect and the setting does not drift into BBM, but a stopper resistance must be inserted after the filter capacitor, and the efficiency is lower. If the overlap time of the contacts is short, the stopper can be quite small, and the amplifier sees a fairly low impedance at all times. It is better, when possible, to use the BBM arrangement, and eliminate the pickup currents, etc., which cause the spikes, since the circuit is then more likely to continue working in the event of partial contact failure. It is still necessary to ensure that the chopper does not drift into the MBB state if

a filter capacitor is present, unless a stopper is included as a precaution.

Transformer-coupled circuits must either be MBB or else contain a buffer capacity (which may be winding self-capacitance) to prevent inductive spikes and a loss of efficiency.

STEADY-STATE EQUIVALENT CIRCUITS. There is a great variety of practical modulator circuits, based on permutations of the principles already discussed, and to analyze their steady-state performance it is first necessary to reduce each to its simplest form. One is then left with a reasonable number of basic equivalent circuits to consider in detail.

Equivalents have already been shown for the DC isolating circuit and the multi-stage output filter. A multiple input filter is equivalent to a single-stage filter with the same series resistance (Figure 13), and two inputs summed at one contact can be reduced to a single equivalent input (Figure 14).

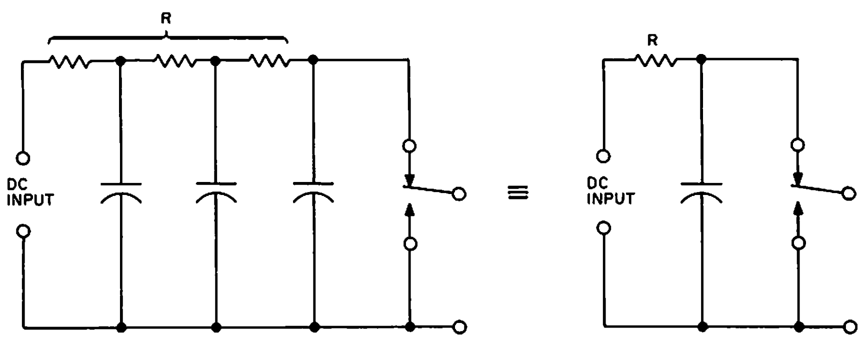


FIGURE 13. *Single and multi-stage input filters are equivalent in the steady state.*

These are fairly obvious points, but a little more care is needed when there are two separate inputs in series. Every modulator has a DC-carrying loop, and the two inputs are always in series in this loop, no matter how the AC signal is taken off. Each input can be represented by a voltage and a source resistance, and the effect of this resistance depends on whether the DC it carries is steady or pulsating, i.e., on whether or not there is a filter on that input. If the current is pulsating, the effect of the resistance depends not only on the mean value of the current, but also on the mark-space ratio of the current pulses. Thus a few rules can be deduced governing how circuits can be manipulated in deriving equivalents. But first we need two definitions.

VOLTAGE GAIN. The voltage gain of a modulator circuit is defined as the peak-to-peak output voltage divided by the DC input voltage.

INPUT RESISTANCE. This is defined as the (steady) input voltage divided by the mean input current.

RULES.

1. A given input voltage, injected in any part of the DC loop, always produces the same AC output. (Otherwise two equal and opposite signals would not give zero output.) In other words, the circuit gain, referred to any input position, will always be the same for a particular circuit.
2. No matter where a given input voltage is injected, the mean DC circulated in the DC loop is always the same for a particular circuit. In other words the input resistance is the same at all points in series with the DC loop.
3. If two separate inputs are both filtered so that each source resistance carries steady DC, they can be considered as a single filtered input (Figure 15).

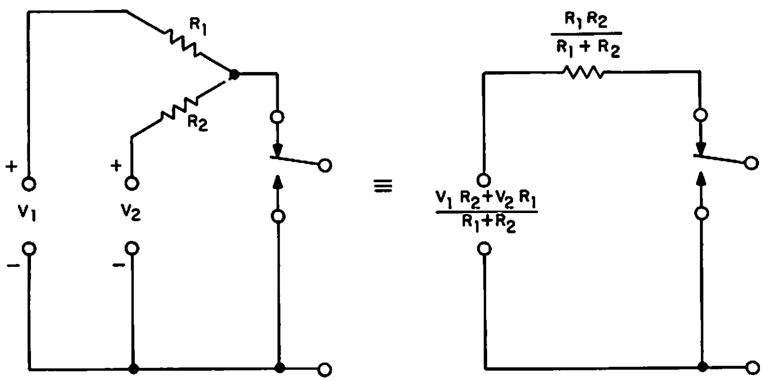


FIGURE 14. *Inputs summed on one contact reduce to a single input.*

4. If one input is filtered and the other is not, the circuit cannot be rearranged in this manner unless (as quite often occurs) one of the source resistances is negligible.
5. If neither input is filtered, the circuit cannot be rearranged unless the current pulses through the two source resistances have the same mark-space ratio. In general terms they do not, although in specific circuits they often may.

We can now list the basic capacity-coupled circuits which are needed to describe the steady-state operation of a very wide range of practical circuits. It turns out that there are six. There are three common switching arrangements:

1. On-off.
2. SPDT, BBM.
3. SPDT, MBB.

Both SPDT circuits have the moving contact connected to the output. One can envisage a two-pole changeover circuit as well, but it would seldom

be practicable on account of grounding problems. Each of these three arrangements can be used in circuits with:

1. Two unfiltered inputs.
2. One filtered and one unfiltered input.

If both inputs are filtered, they can be lumped together as one. These six basic circuits are analyzed in the next section. In each case it will be assumed that there are two inputs. In the on-off and SPDT, BBM circuits, each input will be assumed to have finite source resistance, the resulting expressions thus being as general as possible. It is easy to simplify them for the case where one source resistance vanishes, as often occurs in practice. In the SPDT, MBB case, one of the source resistances will perforce be assumed zero, since otherwise the arithmetic and the resulting expressions become rather heavy.

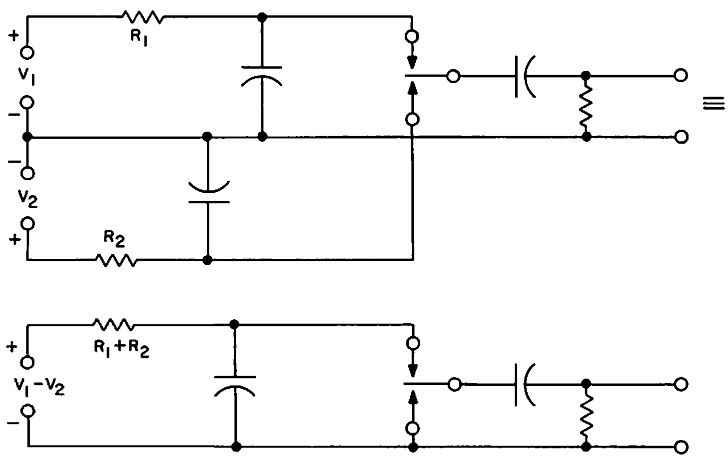


FIGURE 15. *Two filtered inputs can be regarded as a single filtered input.*

We can also list the basic transformer-coupled circuits. On-off versions are both inefficient and subject to pickup; if one takes the trouble to build in a transformer, it is worthwhile using a balanced arrangement. The possible switching arrangements are thus:

1. SPDT, BBM.
2. Two-pole changeover, BBM.
3. SPDT, MBB.
4. Two-pole changeover, MBB.

It is not feasible to feed a transformer from a high resistance source, so filtered inputs are assumed. If a feedback input is to be injected after the filter (a common arrangement), its source resistance will be assumed negligible (as it almost invariably is). Transformer-coupled circuits will be dealt with later.

SECTION III

ANALYSIS OF CAPACITY-COUPLED MODULATORS

Expressions for gain and input resistance of the six basic circuits are listed in Tables 1, 2 and 3, together with the value of stopper resistance (where one exists) giving greatest gain, the optimized gain expression, and the corresponding input resistance. Terminology is given on figures in the tables, x and y being contact closure times, and z, the common time expressed as fractions of a cycle. The method of analysis is similar for each circuit, and the following examples are typical.

ON-OFF MODULATOR WITHOUT FILTER. Consider the on-off modulator with two unfiltered inputs, and for simplicity assume that input voltage No. 2 is zero (Figure 16). Both source resistances are retained, and there is no loss of generality. If we find the gain and input resistance for input No. 1, the results will apply also to input No. 2.

The coupling capacitor is assumed sufficiently large to pass on square waves without appreciable droop. It therefore assumes a steady potential, V_2, after a few initial cycles, and can be regarded as a battery. The circuit exists in one or other of two circuit states (Figure 16) during which steady current i_1, i_2, and i_3 flow through the resistors, being given by the following expressions:

$$V_1 = i_2(R_1 + R_2) + i_3 R_1$$
$$V_2 = i_2 R_2 - i_3 R_3$$
$$V_2 = i_1(R_2 + R_3)$$

V_1 being the input voltage.

Since no charge accumulates on the coupling capacitor, we have a fourth expression:

$$xi_3 = (1 - x)i_1$$

which, with the other three, permits the currents to be expressed in terms of V_1 and the resistances. The gain of the circuit is the peak-to-peak output voltage divided by the input voltage, i.e.,

$$\text{Gain, } A = \frac{(i_1 + i_3)R_3}{V_1}$$

which, with the values of i_1 and i_3 inserted, becomes:

$$A = \frac{1}{\left[1 + \dfrac{R_1}{R_2} + \dfrac{xR_2}{R_3} + \dfrac{R_1}{R_3}\right]}$$

The input resistance is the input voltage V_1, divided by the mean input current, i.e.,

$$\text{Input Resistance, } R_{in} = \frac{V_1}{(i_2 + i_3)x}$$

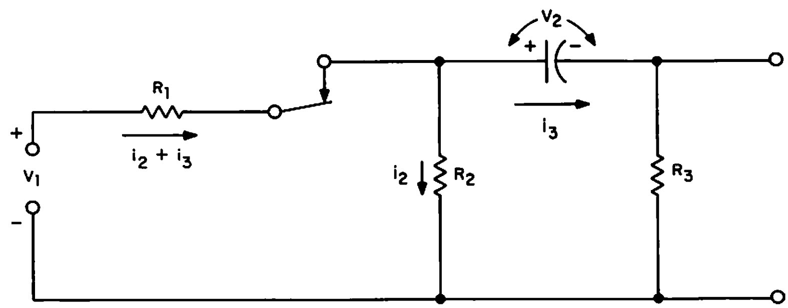
which becomes:

$$R_{in} = \frac{R_1}{x} + R_2 + \left(\frac{1-x}{x}\right)\left(\frac{R_2 R_3}{R_2 + R_3}\right)$$

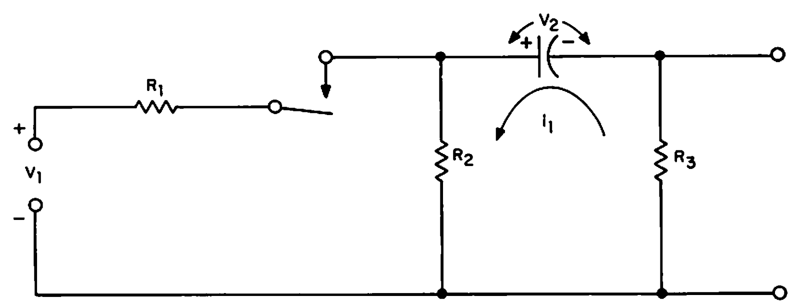
These expressions hold for both inputs. For greatest gain, R_1 should

be as low as possible, and R_2 should have the optimum value, R_2 (opt) $= \sqrt{R_1 R_3 / x}$, obtained by differentiating the expression for gain with respect to R_2.

Inspection of the expression for gain shows that if there is only one source having appreciable resistance, it is better to make this No. 2 input, in which case R_1 can be made zero and R_2, which should then be as low as possible, is the source resistance in question. In other words, the shunt-switching arrangement is better than the series-switching arrangement. If, for some reason, the only source having appreciable resistance has to be made No. 1 input, then R_2 is a circuit resistor and should be given its optimum value. If there are two sources, each with appreciable resistance, then R_1 should be the lower of the two if the only consideration is obtaining the greatest possible gain. In practice, stability considerations will probably decide which input is which.



a. SWITCH CLOSED FOR X



b. SWITCH OPEN FOR 1-X

FIGURE 16. *On-off modulator. Circuit states.*

ON-OFF MODULATOR WITH FILTER. It is sufficient to consider the circuit with one input filtered and the other not filtered. Should both be filtered, they can be lumped together and considered as a single input. It does not matter which input we consider, so let it be input No. 1.

Assume then, that a filter capacitor is connected across input No. 1 after the source resistance R_1, which becomes part of the filter. There may

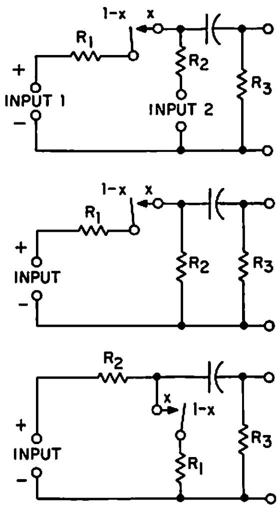
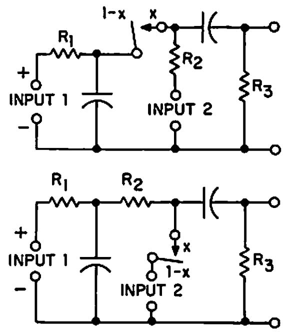
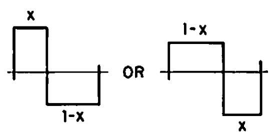
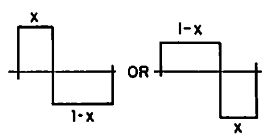
	WITHOUT FILTER	WITH FILTER
CIRCUIT AND SOME EQUIVALENTS		
WAVEFORM		
VOLTAGE GAIN	$\dfrac{1}{1+\dfrac{R_1}{R_2}+x\dfrac{R_2}{R_3}+\dfrac{R_1}{R_3}}$	$\dfrac{1}{1+x\left[\dfrac{R_1}{R_2}+\dfrac{R_2}{R_3}+\dfrac{R_1}{R_3}\right]}$
INPUT RESISTANCE	$\dfrac{R_1}{x}+R_2+\left(\dfrac{1-x}{x}\right)\left(\dfrac{R_2 R_3}{R_2+R_3}\right)$ (MEAN VALUE)	$R_1+R_2+\left(\dfrac{1-x}{x}\right)\left(\dfrac{R_2 R_3}{R_2+R_3}\right)$
R_2 (opt) FOR MAXIMUM GAIN	$\sqrt{\dfrac{R_1 R_3}{x}}$	$\sqrt{R_1 R_3}$
MAXIMUM GAIN	$\dfrac{1}{1+2\sqrt{\dfrac{xR_1}{R_3}}+\dfrac{R_1}{R_3}}$	$\dfrac{1}{1+2x\sqrt{\dfrac{R_1}{R_3}}+x\dfrac{R_1}{R_3}}$

TABLE 1. *Properties of on-off circuits.*

in fact, be separate source and filter resistances, but it makes no difference. The circuit is depicted in Table 1. The gain and input resistance from this capacitor onwards are obtained by writing $R_1 = 0$ in the expressions already obtained.

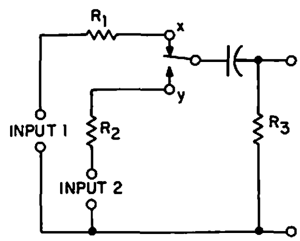
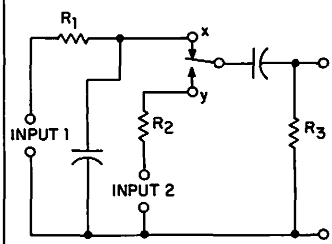
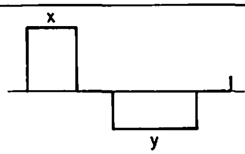
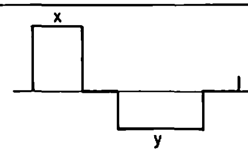
	WITHOUT FILTER	WITH FILTER
CIRCUIT		
WAVEFORM		
VOLTAGE GAIN	$\dfrac{1}{1+\left(\dfrac{xy}{x+y}\right)\left[\dfrac{1}{x}\dfrac{R_1}{R_3}+\dfrac{1}{y}\dfrac{R_2}{R_3}\right]}$	$\dfrac{1}{1+\left(\dfrac{xy}{x+y}\right)\left[\dfrac{R_1}{R_3}+\dfrac{1}{y}\dfrac{R_2}{R_3}\right]}$
INPUT RESISTANCE	$\dfrac{R_1}{x}+\dfrac{R_2}{y}+\left(\dfrac{x+y}{xy}\right)R_3$	$R_1+\dfrac{R_2}{y}+\left(\dfrac{x+y}{xy}\right)R_3$

TABLE 2. *Properties of SPDT, BBM circuits.*

The gain from input No. 1 is therefore:

$$\text{Gain} = A_{(R_1 = 0)} \times \frac{R_{in\,(R_1 = 0)}}{R_1 + R_{in\,(R_1 = 0)}} = \frac{1}{1 + x\left[\dfrac{R_1}{R_2} + \dfrac{R_2}{R_3} + \dfrac{R_1}{R_3}\right]}$$

And the input resistance from the same point is:

$$\text{Input resistance} = R_1 + R_{in(R_1 = 0)} = R_1 + R_2 + \left(\frac{1-x}{x}\right)\left(\frac{R_2 R_3}{R_2 + R_3}\right)$$

Again these expressions apply for both inputs. Again R_1 should be as low as possible, and preferably zero, while R_2 should have an optimum value which in this case is R_2 (opt) $= \sqrt{R_1 R_3}$. The circuit can be arranged

in two ways, as shown in Table 1, by reversing the positions of R_2 and the switch. These are clearly identical, since the alternating voltage across R_2 must be the same as that across the switch.

COMPARISON OF CIRCUIT EFFICIENCIES. As we have already seen, there are many factors which govern the choice of circuit type, and

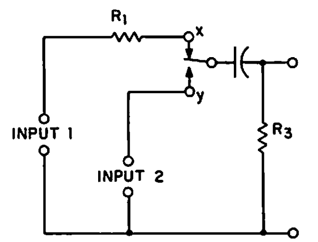
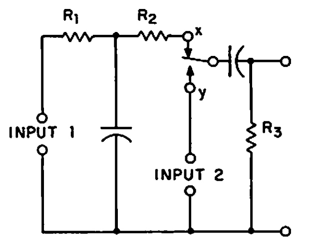
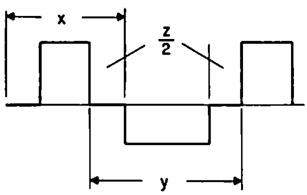
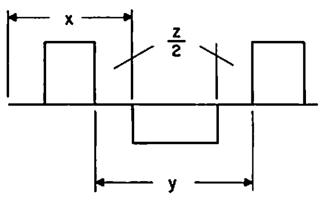
	WITHOUT FILTER	WITH FILTER
CIRCUIT		
WAVEFORM		
VOLTAGE GAIN	$\dfrac{1}{1 + y\dfrac{R_1}{R_3}}$	$\dfrac{1}{1 + z\dfrac{R_1}{R_2} + y\dfrac{R_2}{R_3} + xy\dfrac{R_1}{R_2}}$
INPUT RESISTANCE	$\dfrac{R_3 + yR_1}{z\dfrac{R_3}{R_1} + xy}$	$R_1 + \dfrac{R_3 + yR_2}{z\dfrac{R_3}{R_2} + xy}$
R_2 (opt) FOR MAXIMUM GAIN		$\sqrt{\dfrac{z}{y}R_1R_3}$
MAXIMUM GAIN		$\dfrac{1}{1 + 2\sqrt{yz\dfrac{R_1}{R_3}} + xy\dfrac{R_1}{R_3}}$

TABLE 3. *Properties of SPDT, MBB circuits.*

efficiency is only one of them. The main value of the expressions listed in Tables 1 to 3 is therefore in assisting the choice of component values, and assessing the performance of an already chosen configuration. Nevertheless it is interesting to compare the circuits on the score of efficiency.

NOTES: $h = R_S/R_3$ R_2 = OPTIMUM WHEN PRESENT

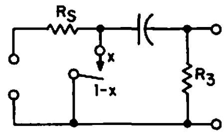
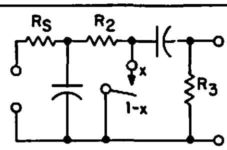
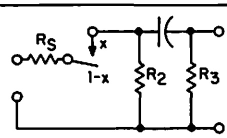
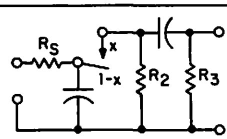
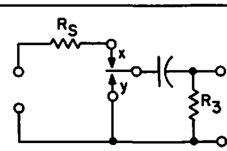
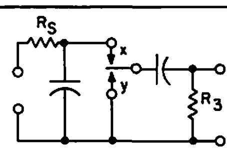
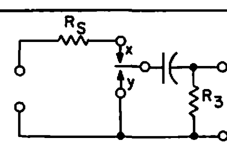
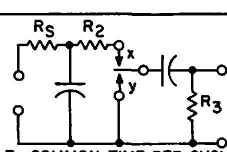
DESCRIPTION	CIRCUIT	VOLTAGE GAIN (h USUALLY < 1)	CURRENT/VOLTAGE GAIN (h USUALLY > 1)
ON-OFF SHUNT		$\frac{1}{1+xh}$	$\frac{1}{R_S}\cdot\frac{1}{\frac{1}{h}+x}$
ON-OFF SHUNT + FILTER		$\frac{1}{1+2x\sqrt{h}+xh}$	$\frac{1}{R_S}\cdot\frac{1}{\frac{1}{h}+2x\sqrt{\frac{1}{h}}+x}$
ON-OFF SERIES		$\frac{1}{1+2\sqrt{xh}+h}$	$\frac{1}{R_S}\cdot\frac{1}{\frac{1}{h}+2\sqrt{\frac{x}{h}}+1}$
ON-OFF SERIES + FILTER		$\frac{1}{1+2x\sqrt{h}+xh}$	$\frac{1}{R_S}\cdot\frac{1}{\frac{1}{h}+2x\sqrt{\frac{1}{h}}+x}$
SPDT, BBM		$\frac{1}{1+\left(\frac{y}{x+y}\right)h}$	$\frac{1}{R_S}\cdot\frac{1}{\frac{1}{h}+\frac{y}{x+y}}$
SPDT, BBM + FILTER		$\frac{1}{1+\left(\frac{xy}{x+y}\right)h}$	$\frac{1}{R_S}\cdot\frac{1}{\frac{1}{h}+\frac{xy}{x+y}}$
SPDT, MBB		$\frac{1}{1+yh}$	$\frac{1}{R_S}\cdot\frac{1}{\frac{1}{h}+y}$
SPDT, MBB + FILTER	 Z= COMMON TIME PER CYCLE	$\frac{1}{1+2\sqrt{yzh}+xyh}$	$\frac{1}{R_S}\cdot\frac{1}{\frac{1}{h}+2\sqrt{\frac{yz}{h}}+xy}$

TABLE 4. *Gains of circuits having only one source resistance.*

To simplify matters, we shall assume that one of the two source resistances is negligible, as is almost always the case in practice. The circuits can then be compared on the basis of the gain which each provides for a given ratio h, of source to load resistance. (R_2, when present, is assumed to have its optimum value.) It makes no difference whether we compare voltage gains, or current/voltage gains, i.e., peak-to-peak load current/input voltage, and both are listed in Table 4.

The basic two-input on-off circuit here divides into two distinct arrangements according to which input position we consider, one having the switch in series with the input chosen, the other having it in shunt across the input. We thus have eight cases, though the two filtered on-off circuits, which look different at first sight, are in fact the same. Bearing in mind that x and y are less than unity, inspection of Table 4 shows:

1. The on-off circuits are in the following descending order of efficiency.
   a. On-off shunt.
   b. On-off shunt with filter, On-off series with filter } equal.
   c. On-off series.
2. The SPDT, BBM circuit is more efficient if its input is filtered. If $x = y = \frac{1}{2}$, it is better than any of the on-off circuits.
3. BBM is better than MBB, but there is not a lot of difference if the common time, z, is small.
4. Addition of a filter capacitor improves the efficiency in all cases where this capacitor is not intermittently shorted. It still does so in the SPDT, MBB circuit if the common time is short.

WAVEFORM SYMMETRY. All the on-off circuits provide more voltage gain if the switch closure time, x, is made small. This is not very significant however, since h is usually small in cases where the criterion is voltage gain, and this accordingly usually approaches unity.

If the modulator works into a low resistance, however, such as a transistor amplifier, h may well be high, and we are concerned with how much current is delivered to the load, not with the voltage developed across it. Expressions for current/voltage gain are given in the last column of Table 4, and it can be seen that, if h is large, it may be worthwhile making x small in the on-off circuits (and y small in the others). In the limit, if the load is negligible, the current delivered to it is inversely proportional to x (or y). For this reason, asymmetrical switching is sometimes used when the load is low and current output is the criterion. There is a disadvantage, however, in that the demodulator must be able to supply more current during the shortened part of the waveform if it is to provide the same DC output.

SECTION IV

ANALYSIS OF MBB TRANSFORMER-COUPLED MODULATORS

In this section we shall consider two very similar transformer-coupled circuits, both having contact actions which ensure that the transformer primary is never open-circuited or presented with a high impedance. This eliminates the spikes which otherwise would occur unless the transformer were buffered by capacity. Both circuit arrangements are symmetrical ones, and in both cases we assume symmetrical switching as well, since asymmetry would give no advantage.

One of the circuits uses a center-tapped primary together with an SPDT, MBB chopper, while the other has a single primary and a two-pole changeover chopper; the latter may in practice be two SPDT, MBB units operated synchronously. The SPDT arrangement is the more common, since the chopper is simpler; additional complexity in the transformer is of less significance. The changeover circuit has the moderate advantage of eliminating the effect of any thermal e.m.f.'s generated within the transformer, but these can be kept small by suitable design.

From the analytical point of view, the only difference between the two arrangements is that the changeover circuit may contain a brief period (t_4) during each changeover action, when the source is not shorted out. This may occur if the two choppers are not well enough synchronized, and does not matter very much in practice, since bad synchronism never by itself open circuits the primary. In the SPDT circuit, the source is either connected to one of the half primaries or else shorted out, and t_4 does not exist.

The changeover circuit is therefore slightly more general, and is used as the basis for an analysis in which no major approximations are made. This gives rise to rather unwieldy expressions which are simplified according to suitable approximations, one of which is that $t_4 = 0$.

TERMINOLOGY.

$B = R_e t_2/L$.
i = Instantaneous magnetizing current at time t, amperes.
$i_a = i + i_b$, amperes.
i_b = Instantaneous current in load, referred to primary, amperes.
$-i_o$ = Value of i at $t = 0$, amperes.
$+i_o$ = Value of i at $t = t_2$, amperes.
I = Mean current drawn from C_1, amperes.
k = Secondary/primary turns ratio, changeover circuit,
= Secondary/half primary turns ratio, SPDT circuit.
L = Inductance of primary winding, changeover circuit, henrys,
= Inductance of half primary winding, SPDT circuit, henrys.

TERMINOLOGY (Cont'd)

Q_2 = Net charge drawn from C_1 during t_2, coulombs.
R_1 = Filter resistance, ohms.
R_2 = Stopper resistance, ohms.
R_3 = Load resistance, ohms.
R'_3 = R_3/k^2, ohms.
R_s = Secondary winding resistance, ohms.
R'_s = R_s/k^2, ohms.
R_p = Resistance of primary winding, changeover circuit, ohms.
$4R_p$ = Resistance of primary winding, SPDT circuit, ohms.
R_a = $R_2 + R_p$, ohms.
R_b = $R'_3 + R'_s$, ohms.
R_o = Input resistance, measured at C_1, ohms.
R_e = $R_a R_b/(R_a + R_b)$, ohms.
t_1 = Cycle time, second.
$2t_2$ = Pulse time per cycle, second.
$4t_3$ = Common time per cycle, second.
$2t_4$ = Open time per cycle, changeover circuit, second,
= Zero, SPDT circuit.
V = Direct input voltage, volts.
V_1 = Voltage across C_1, volts.
V_3 = Twice mean output voltage during t_2, volts.

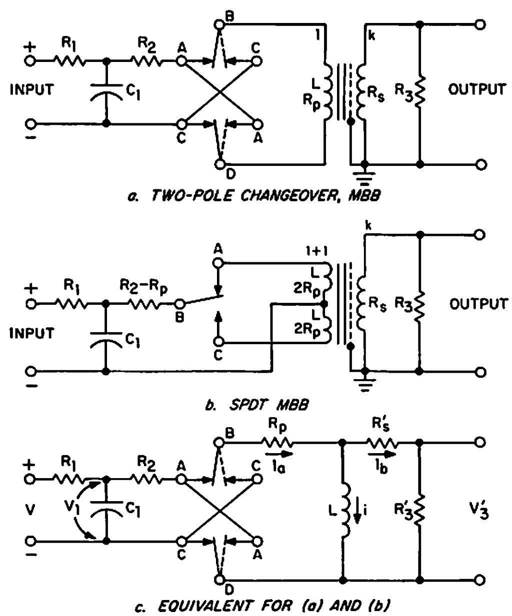


FIGURE 17. *MBB transformer — coupled modulators.*

THE GENERAL CASE. The two practical circuits and their single equivalent circuit are shown in Figure 17. Both need an input filter so that the transformer is fed from a low-impedance source and the waveform is as square as possible. Both need a stopper resistance to control the current drain from the filter capacitor during switchover. The changeover switch is assumed to be made up of two MBB, SPDT choppers, each having a common time per cycle of $2t_3$, and slightly out of synchronism. This is equivalent to a common time per cycle of $4t_3$ in the SPDT circuit. When all three contacts of the SPDT circuit are joined, the shorted primary

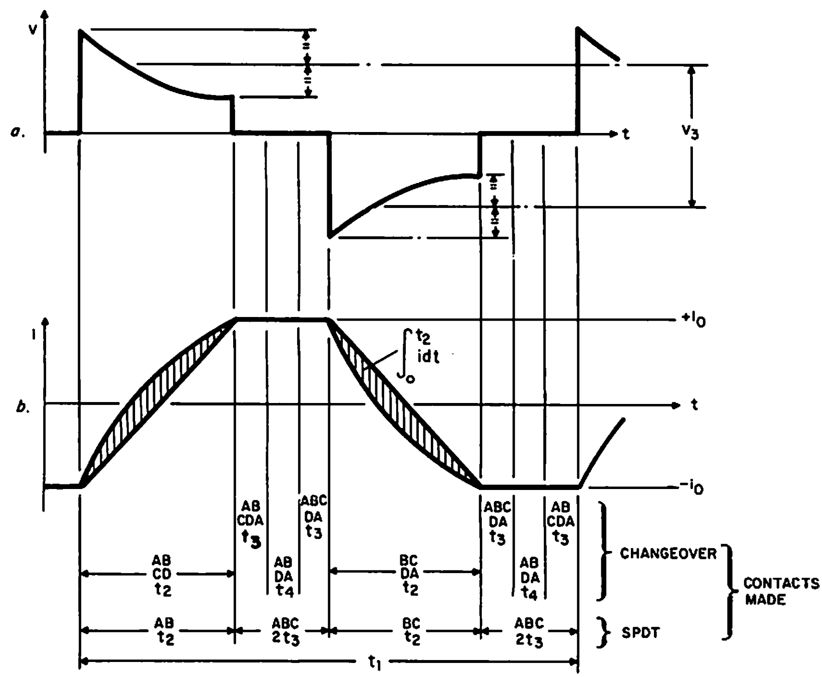


FIGURE 18. *Waveforms of output voltage and magnetizing current.*

looks to the filter like a resistance R_p, since equal and opposite currents flow in the two half primaries, and induce no net flux or back e.m.f. The stopper resistor is therefore labelled $R_2 - R_p$, so that the single equivalent circuit is true for both cases. Transformer leakage reactance and self capacitance are neglected, but losses can be accounted for if necessary by assuming the load resistance to include the transformer loss resistance.

The filter capacitance is assumed large, so that steady voltage exists across it when steady-state conditions have been reached after a number of initial cycles. Figure 18, shows the typical output voltage waveform in which both the droop and the common time have been exaggerated for

clarity. The magnetizing current waveform is also shown. This can be observed practically on a changeover circuit with the load removed. In the SPDT arrangement, the primaries carry the filter short-out current in addition, as shown in Figure 19.

Operation of the SPDT circuit is considered in rather more detail later, and shown to be essentially the same as that of the equivalent circuit. Consider therefore the equivalent circuit, Figure 17c. At time $t = 0$, the filter capacitor is connected via R_2 to the transformer, and currents flow in the circuit, whose instantaneous values are i_a, i and i_b. After a

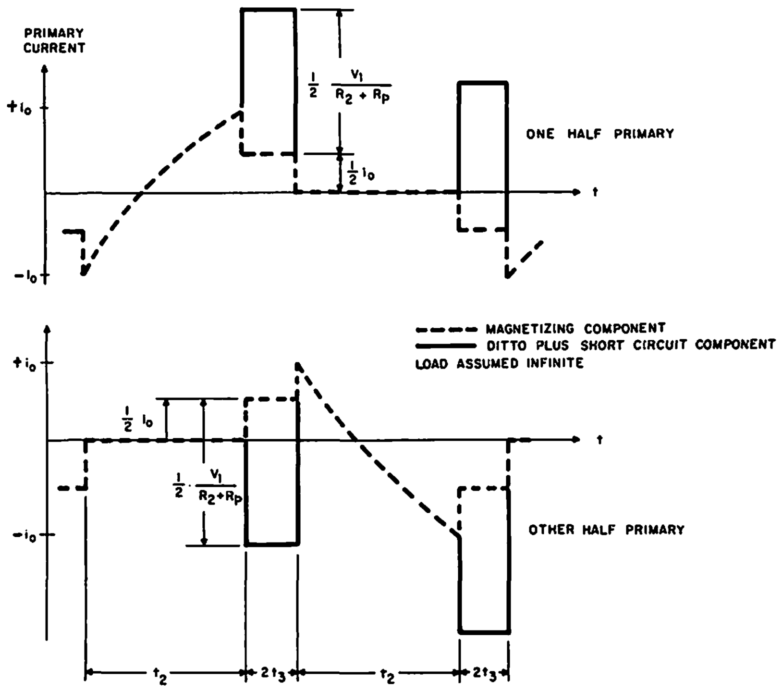


FIGURE 19. *Primary current waveforms. SPDT case.*

time t_2, the current in L has risen to a value $+ i_o$ and the transformer is short-circuited for a time $t_4 + 2t_3$. During this short-circuit period, the flux in the transformer core is maintained by i_o which remains sensibly constant since the primary self time constant, L/R_p is relatively long.

The next half cycle therefore commences with a current $+ i_o$ still flowing through L and tending to charge the capacitor, now connected the other way round, back to its original potential. The current falls during t_2 to a final value which, by symmetry, must be $-i_o$, and there is then another short circuit period, after which the cycle begins again. Because of the

symmetry, it is sufficient to analyze a single half cycle. This gives expressions for the currents, the mean drain from the capacitor, the input resistance and the gain.

Let
$$R_a = R_2 + R_p$$
$$R_b = R'_s + R'_3$$
$$R_e = \frac{R_a R_b}{R_a + R_b}$$

Then the basic circuit equations for the period t_2 are:

$$V_1 = R_a i_a + L\, \delta i/\delta t$$
$$L\, \delta i/\delta t = R_b i_b$$
$$i_a = i + i_b$$

Eliminating i_a and i_b from the above, integrating, and putting $i = -i_o$ when $t = 0$, and $i = +i_o$ when $t = t_2$, gives:

$$i = \frac{V_1}{R_a}\left(1 - \frac{2\epsilon^{-R_e t/L}}{1 + \epsilon^{-R_e t/L}}\right)$$

This enables us to find the charge drawn from C_1 during t_2, which is:

$$Q_2 = \int_0^{t_2} (i + i_b)\,\delta t = \int_0^{t_2} i\delta t + \frac{L}{R_b}\Big[\, i \,\Big]_0^{t_2}$$
$$= \frac{V_1}{R_a}\left[t_2 - 2\frac{L}{R_a}\left\{\frac{1 - \epsilon^{-R_e t_2/L}}{1 + \epsilon^{-R_e t_2/L}}\right\}\right]$$
$$= \frac{V_1}{R_a}\left[t_2 - 2\frac{L}{R_a}\, tanh\, \frac{1}{2}\,\frac{R_e t_2}{L}\right]$$

For convenience, the *tanh* term is now replaced by the first two terms of its expansion as a power series. Provided that $R_e t_2/L$ is not greater than unity, the consequent error in Q_2 is below 1%, and we are not interested in greater values, since they imply a peaky waveform whose droop in fact exceeds 63%. Thus:

$$Q_2 = \frac{V_1}{R_a}\left[\frac{R_a t_2}{R_a + R_b} + \frac{1}{12}\,\frac{R_e^3\, t_2^3}{R_a L^2}\right]$$

Now the charge drawn from C_1 during a complete half cycle is:

$$Q_2 + 2t_3 V_1/R_2$$

the second term representing the short-circuiting of C_1 through R_2 during the switchover action, and the mean current drain from C_1 is therefore:

$$I = \frac{2}{t_1}\left[Q_2 + 2t_3\, V_1/R_2\right]$$

The input resistance of the circuit, measured at C_1 is thus:

$$R_o = \frac{V_1}{I} = \frac{R_a + R_b}{\frac{2}{t_1}\left[2t_3\left(\frac{R_a + R_b}{R_2}\right) + t_2 + \frac{1}{12}\ R_e^2\ \frac{R_b\, t_2^3}{R_a\, L^2}\right]}$$

The input resistance, measured at the input terminals of the circuit, is of course obtained by adding R_1 to the above value of R_o.

We can now find the gain. The voltage across L at any time during t_2 is:

$$L\frac{\delta i}{\delta t} = 2V_1\frac{R_e}{R_a}\left[\frac{\epsilon^{-R_e t/L}}{1 + \epsilon^{-R_e t_2/L}}\right]$$

and the sum of its values at $t = 0$ and $t = t_2$ is:

$$2V_1\ R_e/R_a$$

Multiplying this by R'_3/R_b, we obtain the output across R'_3, namely:

$$2\,V_1\ R_e\ R'_3/R_aR_b$$

and the gain, from the filter capacitor onwards, is:

$$2k\ R'_3/(R_a + R_b)$$

where k is the turns ratio of the transformer. The gain from the input terminals is obtained by multiplying this expression by $R_o/(R_1 + R_o)$ and is:

$$\text{Gain} = \frac{2\,k}{\frac{R_a + R_b}{R'_3} + \frac{2\,R_1}{t_1\,R'_3}\left[2t_3\left(\frac{R_a + R_b}{R_2}\right) + t_2 + \frac{1}{12}\ \frac{R_b}{R_a}\ \frac{R_e^2}{L^2}\ t_2^3\right]}$$

This has its maximum value, with R_2 varying when:

$$2t_3\left(\frac{R_b + R_p}{R_2^2}\right) = \frac{1}{2}\ \frac{t_1}{R_1} + \frac{1}{12}\ \frac{t_2^3}{L^2}\ (R_b - R_a)\left(\frac{R_b}{R_a + R_b}\right)^3$$

APPROXIMATIONS. The expression for gain just developed is accurate to 1% provided only that C_1 is very large and $R_e t_2/L$ is not greater than unity. Beautiful though it may be in its generality, it is not of very much use as it stands. But for those who want to make their own deductions, it may be useful.

It seems that there are several ways of deriving simpler approximate expressions, and we shall start by making the following assumptions:

1. Neglect the winding resistances. This gives little error in most practical cases, unless one is using a transformer with more inductance than necessary, and hence too much winding resistance. Having designed a circuit on this assumption, it is simple to check that, in fact, $R_p \ll R'_3$, and $R_s \ll R_3$.

2. Put $t_4 = 0$, so that in consequence $2t_2 + 4t_3 = t_1$. This is always true of the SPDT circuit, and the changeover circuit would always be set up so that it were true initially.

3. Write R'_3 as R_3/k^2 so that k appears explicitly in the expression.

4. Write $R_c t_2/L$ as B, a quantity which we have already assumed to be less than or equal to unity.

Assuming the foregoing, and no more, we can simplify the expression for gain to:

$$\text{Gain} = \frac{2}{k\left[\dfrac{R_2}{R_3} + \dfrac{R_1}{R_3}\right] + \dfrac{1}{k}\left[1 + 4\dfrac{t_3}{t_1}\dfrac{R_1}{R_2}\right] + \dfrac{1}{6k}\dfrac{R_1}{R_2}\dfrac{t_2}{t_1}B^2}$$

To simplify matters still further, and because L can often be made large, we next assume L to be infinite. (Just how large L has to be to make this legitimate can be deduced very easily, but the resulting expression is not very informative.) We then obtain:

$$\text{Gain } (L \text{ infinite}) = \frac{2}{k\left[\dfrac{R_2}{R_3} + \dfrac{R_1}{R_3}\right] + \dfrac{1}{k}\left[1 + 4\dfrac{t_3}{t_1}\dfrac{R_1}{R_2}\right]}$$

There is clearly a value of R_2 which makes this a maximum. Since R_2 is merely a circuit resistor, we are interested in this value which is:

$$R_2 \text{ (opt)} = \sqrt{4\frac{t_3}{t_1}\frac{R_1 R_3}{k^2}} \quad \text{(for any value of } k\text{)}$$

Similarly, there is a value of transformer ratio, k, which makes the gain a maximum. If one is designing a circuit, this too is under control, and we are interested in it.

$$k \text{ (opt)} = \sqrt{\left(\frac{R_3}{R_1 + R_2}\right)\left(1 + 4\frac{t_3}{t_1}\frac{R_1}{R_2}\right)} \quad \text{(for any value of } R_2\text{)}$$

If R_2 and k are both optimized at the same time, we get:

$$R_2 \text{ (opt opt)} = R_1\sqrt{4\frac{t_3}{t_1}}$$

$$k \text{ (opt opt)} = \sqrt{\frac{R_3}{R_1}}$$

The last expression is the same as obtains for correct matching in sine wave circuits. Putting these simultaneous optimum values into the expression for gain (L not infinite) we obtain the optimum gain:

Gain (opt R_2, opt k) $=$

$$\left(\frac{1}{1+\sqrt{4\dfrac{t_3}{t_1}}}\right)\sqrt{\frac{R_3}{R_1}}\left[\frac{1}{1+\dfrac{1}{12}\left(\dfrac{1}{4\dfrac{t_3}{t_1}+\sqrt{4\dfrac{t_3}{t_1}}}\right)\dfrac{t_2}{t_1}B^2}\right]$$

$$= \frac{1}{1+\sqrt{4\dfrac{t_3}{t_1}}}\sqrt{\frac{R_3}{R_1}} \text{ approximately,}$$

provided that $B \leqslant 1$ and $\sqrt{4\dfrac{t_3}{t_1}} \gg \dfrac{1}{24}$

The first condition has been assumed all along, and is necessary in any case if the waveform is to be a useful shape. The second requires a factor of inequality of e.g., 10 for 10% accuracy in the approximate optimum gain expression, or of 100 for 1% accuracy and so on. $4t_3/t_1$ is the fractional common time of the contacts, and might typically be 1/10, giving a factor of inequality of about 7.5, and accuracy of at least 15% in the approximate expression. Most of the above expressions are listed in Table 5, together with the assumptions on which they are based.

SIGNIFICANT PRIMARY RESISTANCE. Sometimes the resistance of the transformer primary is not entirely negligible. If a circuit is designed on the basis of expressions (Table 5) which neglect winding resistance altogether, and a practical transformer has primary resistance R_p, which is significant in relation to R_3/k^2, the following rather more accurate expressions may be of use. The assumptions are:

L infinite.

$t_4 = 0$.

R_s negligible (in a step-up transformer it is nearly always much less than R_3).

$R_p \ll (t_1/4t_3)\, R_2$ (i.e. R_p of the same order as R_2 or less).

And the resulting expressions are:

$$\text{Gain} = \frac{2}{k\left[\dfrac{R_a}{R_3}+\dfrac{R_1}{R_3}\right]+\dfrac{1}{k}\left[1+4\dfrac{t_3}{t_1}\dfrac{R_1}{R_2}\right]}$$

$$R_2\text{ (opt)} = \sqrt{4\frac{t_3}{t_1}R_1\left(\frac{R_3}{k^2}+R_p\right)}$$

$$k\text{ (opt)} = \sqrt{\left(\frac{R_3}{R_1+R_a}\right)\left(1+4\frac{t_3}{t_1}\frac{R_1}{R_2}\right)}$$

LOW TRANSFORMER RATIO. If the transformer ratio k is much smaller than the optimum, then R_3/k^2 becomes large compared with the

ASSUMED IN ALL EXPRESSIONS:

- WINDING RESISTANCES NEGLIGIBLE
- $t_4 = 0$ (INHERENT IN SPDT CIRCUIT)
- $B = \dfrac{R_2R_3/k^2}{R_2+R_3/k^2}\,\dfrac{t_2}{L} \leqslant 1$ (FOR DROOP $\leqslant$ 63%)

PROPERTY	EXPRESSION	ADDITIONAL ASSUMPTIONS
GAIN	$\dfrac{2}{k\left[\dfrac{R_2}{R_3}+\dfrac{R_1}{R_3}\right]+\dfrac{1}{k}\left[1+4\dfrac{t_3}{t_1}\dfrac{R_1}{R_2}\right]+\dfrac{1}{6k}\dfrac{t_2}{t_1}\dfrac{R_1}{R_2}B^2}$	NONE
GAIN	$\dfrac{2}{k\left[\dfrac{R_2}{R_3}+\dfrac{R_1}{R_3}\right]+\dfrac{1}{k}\left[1+4\dfrac{t_3}{t_1}\dfrac{R_1}{R_2}\right]}$	L INFINITE
R_2 (opt) FOR MAXIMUM GAIN	$\sqrt{4\dfrac{t_3}{t_1}\dfrac{R_1R_3}{k^2}}$	L INFINITE
k (opt) FOR MAXIMUM GAIN	$\sqrt{\left(\dfrac{R_3}{R_1+R_2}\right)\left(1+4\dfrac{t_3}{t_1}\dfrac{R_1}{R_2}\right)}$	L INFINITE
R_2 (opt opt)	$R_1\sqrt{4\dfrac{t_3}{t_1}}$	R_2 AND k BOTH OPTIMUM L INFINITE
k (opt opt)	$\sqrt{\dfrac{R_3}{R_1}}$	
GAIN (opt opt)	$\left(\dfrac{1}{1+\sqrt{4\dfrac{t_3}{t_1}}}\right)\sqrt{\dfrac{R_3}{R_1}}$	R_2 AND k BOTH OPTIMUM $\sqrt{4\dfrac{t_3}{t_1}} > \dfrac{1}{24}$ NO ADDITIONAL RESTRICTION ON L
TERM DESCRIBING OUTPUT WAVEFORM	ϵ^{-Bt/t_2}	NONE

TABLE 5. *Properties of MBB transformer — coupled modulators.*

optimum for R_2 based on the assumption of infinite L. The condition for this is:

$$\frac{R_3}{k^2} \gg \sqrt{4\,\frac{t_3}{t_1}\,\frac{R_1 R_3}{k^2}}$$

$$\text{i.e.,}\quad k \ll \sqrt{\frac{t_1}{4t_3}\,\frac{R_3}{R_1}}$$

and may occur even with optimum k if the common time is very small. We can then write R_2 for R_e in the gain expression (neglecting winding resistances but not L), and find that:

$$R_2\ (\text{opt}) = \sqrt{\frac{24\ t_3}{\dfrac{t_2^3}{L^2} + 6\,\dfrac{t_1 k^2}{R_1 R_3}}}$$

which may depend on L. If, in addition, k or L are so low that:

$$k^2 \ll \frac{1}{6}\,\frac{t_2^3}{t_1}\,\frac{R_1 R_3}{L^2}$$

$$\text{then } R_2\ (\text{opt}) = \frac{L}{t_2}\sqrt{\frac{24\ t_3}{t_2}}$$

In this case, since $R_2 t_2/L$ must be less than or equal to unity, we have:

$$24\,\frac{t_3}{t_2} \leqslant 1$$

as a further condition, i.e., the common time must be less than about 8% of the cycle. These expressions may be of use in the restricted case of improper matching (k too low) or of unusually short common time together with low inductance.

NOTE ON THE SPDT CIRCUIT. The analysis has been made in terms of an equivalent circuit which is, in fact, a changeover arrangement. One can see that this is justified by studying the operation of the SPDT circuit a little more closely. During t_2, the current in half the primary rises from $-i_o$ to $+i_o$, as in the equivalent. Then all three contacts are joined, and no potential can appear across the two ends of the primary. Hence the core flux remains almost constant, being maintained by a current $i_o/2$ which circulates through the whole primary. Superimposed on this are two equal currents which flow in opposite directions through the two half primaries under the influence of the input voltage. These set up no flux and no back e.m.f. so, to the input, the shorted transformer looks like a resistance R_p. These currents represent a drain on the filter and are controlled by the stopper; without it, they would be very large. As soon as the first half of the primary is disconnected, these currents cease, and the current in the second half immediately takes up the value $+i_o$, falling during t_2 to $-i_o$ again. Figure 19, shows the waveforms of magnetizing and of total cur-

rent in both half primaries with infinite load. The above rather intuitive description neglects possible interaction between the two components of current, but can be shown to be correct by detailed analysis.

CAPACITANCE ACROSS THE TRANSFORMER. If the transformer primary (both halves in the SPDT case) is open-circuited even briefly while current is flowing, the flux collapses and large voltage spikes are set up unless some capacitance exists across the winding. Apart from causing a bad waveform, this effect wastes the stored magnetic energy and lowers the circuit efficiency. A small amount of capacitance is therefore desirable, even in MBB circuits, to guard against the possibility of contact bounce or similar minor faults. Sufficient capacitance for this purpose is usually inherent in the transformer.

BBM circuits, of course, may need more buffering than this, and the capacitance should be chosen so that the spike, which is really the first part of what would become a train of dying oscillations, just meets the next horizontal part of the waveform at the end of the switching period. However, these circuits need separate analysis.

Sometimes one sees MBB circuits with sufficient shunt capacitance added to resonate the transformer magnetizing inductance at the operating frequency. This appears to be of no value. The capacitance is charged up during each t_2 period and promptly shorted out again during the switching period, the net effect being to drain current unnecessarily from the input. Capacitance of this order may lower the input resistance considerably. It does not even give appreciable discrimination against pickup which has been chopped into twice the operating frequency.

DESIGN PROCEDURE. The following example outlines the sort of procedure to be adopted in designing an MBB transformer-coupled circuit. Suppose that a SPDT circuit is required to work at 50 CPS from a source resistance of 1000 ohms into a load of 100,000 ohms and deliver a waveform whose droop is about 25%.

The common time must be large enough to guarantee continued MBB action in spite of contact drift, and 10% is a reasonable figure, giving $t_3 = 0.5$ m.s. This determines t_2. Winding resistances are neglected to start with, and R_2 and k are optimized assuming that L is infinite. (Table 5). Thus:

$t_1 = 20$ m.s.
$t_2 = 9$ m.s.
$t_3 = 0.5$ m.s.
$R_1 = 1000$ ohms.
$R_3 = 100{,}000$ ohms.
k (opt opt) $= 10$.
R_2 (opt opt) $= 1000\sqrt{2/20} = 320$ ohms approximately.
$R_3/k^2 = 1000$ ohms.
$R_e = 1000 \times 320/1320 = 240$ ohms approximately.

For 25% droop, B must be about 0.25, and hence L must be $4R_e t_2 = 8.6$ henrys. In practice a value of 10 henrys would be specified, measured with small signals applied. Now $\sqrt{4t_3/t_1} = 0.32$ which exceeds 1/24 by a factor of 7.5, so the optimum gain expression is accurate to 15% if $B = 1$. Since $B = 0.25$, it will be accurate to 1%, and the gain is $10/(1 + 0.32) = 7.5$. The secondary winding resistance of a transformer of this type would be negligible in relation to 100,000 ohms. The stopper resistor is given the value $R_2 - R_p$, and each half primary ($2\,R_p$) must therefore be less than 640 ohms, so that this is possible. Inspection of the circuit, Figure 17, shows that R_p should also be small in relation to R'_3, if expressions derived by neglecting R_p are to be true, so there may be a noticeable drop in gain if R_p exceeds about 100 ohms.

Should the transformer design require R_p to be, for example, 200 ohms, the more accurate expressions which allow for R_p may be used. Values of k and R_2 which simultaneously satisfy the expressions for optimum conditions are found by trial and error to be about 9 and 400 ohms respectively, and the corresponding gain is 7.0.

SECTION V

ANALYSIS OF BBM TRANSFORMER-COUPLED MODULATORS

BBM transformer-coupled modulators nearly always take one or the other of two practical forms, namely the two-pole changeover circuit (Figure 20a) or the SPDT circuit (Figure 20b), both of which are represented by the single equivalent circuit shown in Figure 20c.

Unless the source resistance is exceptionally low, it should be followed by a filter capacitor. This not only provides filtering and an output wave-

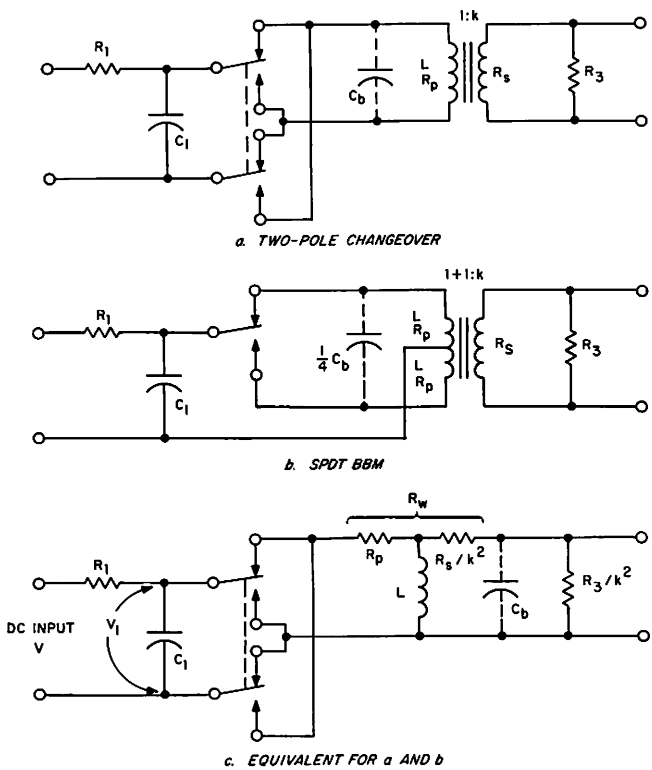


FIGURE 20. *BBM transformer — coupled modulators.*

form having an almost flat top, but also prevents losses which otherwise would be caused by the transformer magnetizing current flowing through the source resistance. The filter is never short circuited, and no stopper resistance is needed between filter and transformer. But a buffer capacitor, C_b, is frequently required to absorb the magnetizing current during the off time, and so control the voltage reversal which then occurs.

In what follows, transformer leakage reactance, self capacitance, and winding resistance are first neglected, as representing small departures from the ideal case. Terminology is shown in Figures 20 and 21, being largely the same as in the MBB case. The following differences are noteworthy:

$2t_2$ = Dwell time per cycle (= Pulse time per cycle), second.
$4t_3$ = Off time per cycle, second.
R_p = Resistance of primary winding, changeover circuit, ohms,
= Resistance of half primary, SPDT circuit, ohms.
$R_w = R_p + R_s/k^2$, ohms.
C_b = Buffer capacitance, referred to primary, mfd.

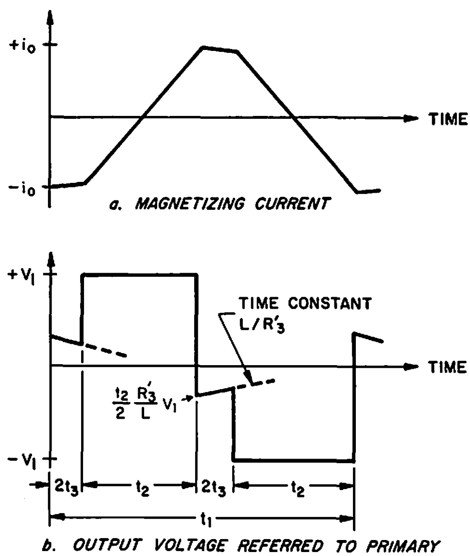


FIGURE 21. *Current and voltage waveforms — large L ($C_b = 0$).*

The mode of operation is rather different according to whether L is large or small, and the two cases (which adjoin) are taken separately.

OPERATION WITH LARGE L. Consider the equivalent circuit shown in Figure 20c. Suppose that L/R'_3 considerably exceeds $2t_3$ and that there is no buffer capacitor.

During one dwell period t_2, the magnetizing current in L rises to a value $+i_o$. The primary is then open circuited, and instantaneously a current i_o flows in the load R'_3. Because of our assumption about L, this current decreases only slightly during the off time, having nearly the same value at the start of the next dwell period, and flowing back into the filter capacitor C_1. Hence the magnetizing current swings between $+i_o$ and $-i_o$ approximately during each dwell period, as shown in Figure 21a. Since

L is large, the currents are small, and there is very little net DC drain from the filter capacitor which, however, prevents AC flowing through the source resistance and so causing I^2R losses.

During the dwell periods, the output voltage referred to the primary, is steady and equal to plus or minus the voltage V_1 on the filter capacitor. Winding resistance may in practice cause some, but as a rule not much, droop. The question is, what happens during the off periods?

The magnetizing current rises at a rate dictated entirely by L and V_1, and its value at the end of the dwell period is:

$$i_o = \frac{1}{2} \frac{t_2}{L} V_1$$

At the instant the primary is open circuited, this current flows in R'_3, giving a reverse voltage:

$$i_o R'_3 = \frac{1}{2} \frac{R'_3}{L} t_2 V_1$$

which decays with time constant L/R'_3 until the next dwell period. For this reverse voltage to be less than or equal to V_1, as is usually required, we have the further condition that:

$$\frac{L}{R'_3} \geqslant \frac{t_2}{2} \qquad \text{(for no spike).}$$

A typical output voltage waveform is shown in Figure 21b. As L is increased, the intermediate steps sink towards the baseline.

This mode of operation is rather uneconomic in that it requires a fairly large value of inductance, but it does have the advantage that no buffering is needed. Although circuits with lower L are just as efficient (neglecting winding resistance) if they are correctly buffered, this condition is not maintained if the contact setting drifts. It is possible to round off the intermediate steps by adding capacity across L, but not possible to make the circuit oscillatory during the off time, since there is little energy associated with L. In practice there is usually some rounding off due to transformer self-capacitance.

GAIN WITH LARGE L. Assuming the off time to be a fairly small fraction of a cycle, the input resistance seen at C_1 is approximately $R'_3 = R_3/k^2$. This is a fair approximation, particularly if the height of the intermediate steps approaches V_1, since a voltage approximating $\pm V_1$ is then maintained across the load all the time. The energy to do this must come from somewhere, and is in fact supplied from C_1 which delivers a net current slightly in excess of V_1/R'_3 during the dwell period.

The voltage across C_1 is therefore:

$$V_1 = V \frac{R_3/k^2}{R_1 + R_3/k^2}$$

and the gain is:

$$\text{Gain} = 2\,k\frac{V_1}{V} = \frac{2\,R_3}{kR_1 + \frac{1}{k} R_3}$$

As k varies, this has a maximum value of:

$$\text{Gain (opt)} = \sqrt{R_3/R_1}$$

$$\text{given by } k \text{ (opt)} = \sqrt{R_3/R_1}$$

Thus correct matching requires the same transformer ratio as would be used in sine wave circuits. The above simple expressions are adequate for most practical cases where L is large.

SIGNIFICANT WINDING RESISTANCE. With large L and many turns, winding resistance may sometimes become significant in relation to the load resistance, causing a drop in gain. A more accurate set of expressions is then:

$$\text{Gain} = \frac{2R_3}{k(R_1 + R_w) + \dfrac{1}{k} R_3}$$

$$k \text{ (opt)} = \sqrt{R_3/(R_1 + R_w)}$$

$$\text{Gain (opt)} = \sqrt{R_3/(R_1 + R_w)}$$

where R_w is the sum of primary resistance (half primary in the SPDT case) and reflected secondary resistance. In most cases, the more approximate value of k (opt) would be used, the gain being checked by the first of the above expressions.

L VERY LARGE. If L is so large that the intermediate waveform steps are close to the baseline, a further refinement can be made in the analysis. Assuming that L is infinite, but allowing for winding resistance, we see that the mean input resistance at C_1 is:

$$\frac{t_1}{2t_2}(R_w + R_3/k^2)$$

$$\text{Hence } V_1 = V \frac{\dfrac{t_1}{2t_2}(R_w + R_3/k^2)}{R_1 + \dfrac{t_1}{2t_2}(R_w + R_3/k^2)}$$

$$\text{And Gain} = 2k\frac{V_1}{V}\,\frac{R_3/k^2}{R_w + R_3/k^2} = \frac{2R_3}{k\left(\dfrac{2t_2}{t_1}R_1 + R_w\right) + \dfrac{1}{k}R_3}$$

$$\text{Whence } k \text{ (opt)} = \sqrt{\frac{R_3}{\dfrac{2t_2}{t_1}R_1 + R_w}}$$

$$\text{And Gain (opt)} = \sqrt{\frac{R_3}{\dfrac{2t_2}{t_1}R_1 + R_w}}$$

The gain, measured in terms of peak-to-peak output, thus increases with increasing off time, but of course the pulse width decreases. Long off

times and the consequent large values of L are generally undesirable and would only be used if there were some special requirement for this type of waveform.

OPERATION WITH SMALLER L. It is more economical to use smaller values of L. As L is reduced, the magnetizing current increases in inverse proportion, and the stored energy $\frac{1}{2}Li^2$ also increases.

Hence the circuit comprising L, R'_3 and a buffer capacitor C_b can be rendered oscillatory during the off periods. If the buffer capacitance is chosen correctly, the oscillation can be made such that the voltage exactly reaches $-V_1$ at the end of the off period. It then precisely matches the voltage on the filter capacitor when this is re-connected in the reverse

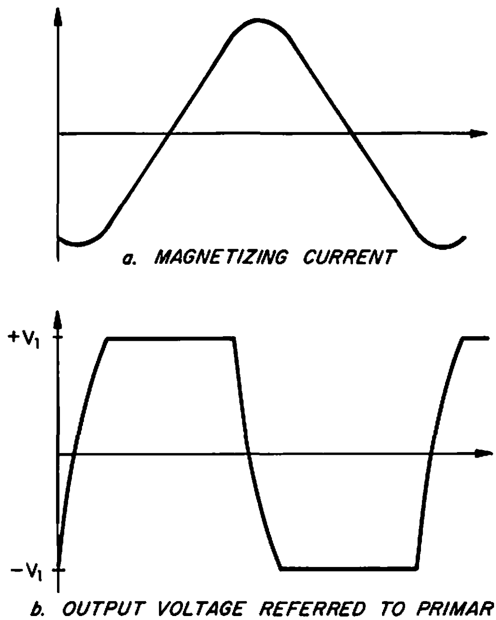


FIGURE 22. *Current and voltage waveforms — low L. Ideal buffering.*

direction. There are no losses due to transients, and the arrangement is analogous to the resonating of an inductance by parallel capacitance in sine wave circuits. Typical current and voltage waveforms are shown in Figures 22a and 22b.

If the off time is not too long, the input resistance is again approximately R'_3, and the optimum values of k and of gain are $\sqrt{R_3/R_1}$ as before. This is quite a good approximation, since an appreciable output is provided during the off time, and corresponding energy is drawn from the source. With lower values of L, winding resistances are in practice lower, and can usually be neglected in relation to the load.

The effects of under buffering (C_b too small) and of over buffering (C_b too large) are shown in Figures 23a and 23b. Both involve a loss of energy which can be shown to be proportional to the buffering error and the nominal value of C_b. In the sine wave analogy, the current drawn by an incorrectly resonated inductance would be proportional to the error and the nominal capacitance.

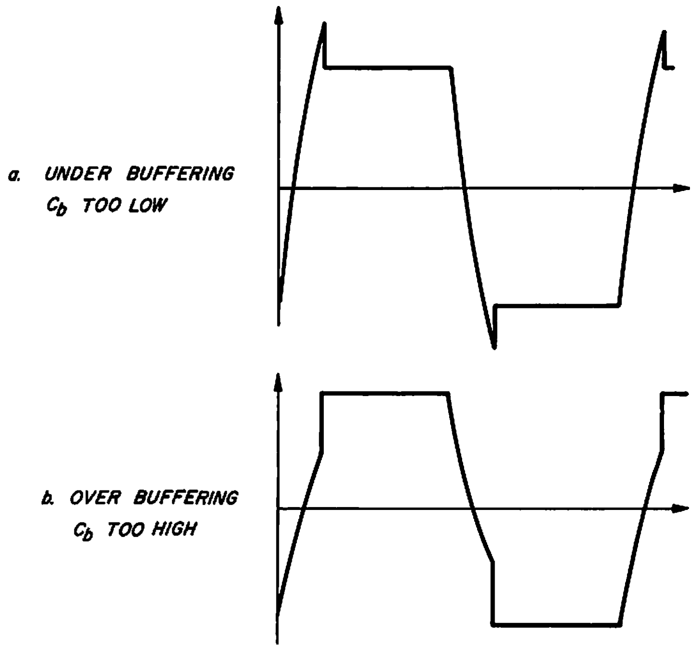


FIGURE 23. *Voltage waveforms — incorrect buffering.*

MAGNITUDE OF L AND C_b. A complete analysis of the circuit has been given by Evans* who shows that it is possible to choose an ideal buffer capacitor, giving just the right amount of reverse swing, provided that:

$$\frac{L}{R'_3} \leqslant \frac{t_1}{5} \text{ to } \frac{t_1}{10} \text{ approximately,}$$

for values of off time typically found. With very small off times, somewhat larger values of L are permissible. The above condition covers off times up to about 30% of the cycle. We may note that this condition takes over approximately where that for the "high L" case with no spikes left off, since $t_1/5$ is roughly equal to $t_2/2$.

The ideal value of buffer capacity is given by:

$$\frac{1}{2}\, t_2 \sqrt{\frac{1}{LC_b}} = \cot\left\{ t_3 \sqrt{\frac{1}{LC_b}} \right\}$$

*Evans R. H., Vibratory Power Converters. Proc. I.E.E. Monograph No. 109 R. September, 1954 (102C, p 62).

which becomes exact as L tends to zero, and is correct within about 10% provided that L/R'_3 is not greater than about $t_1/10$. This expression is depicted as a graph in Figure 24, and can be used to estimate the required value of C_b. This should always be checked experimentally, however, since transformer self capacity may be present. This is especially important if the off time is small, and L is near its upper permitted value, since both these factors tend to make the ideal C_b small.

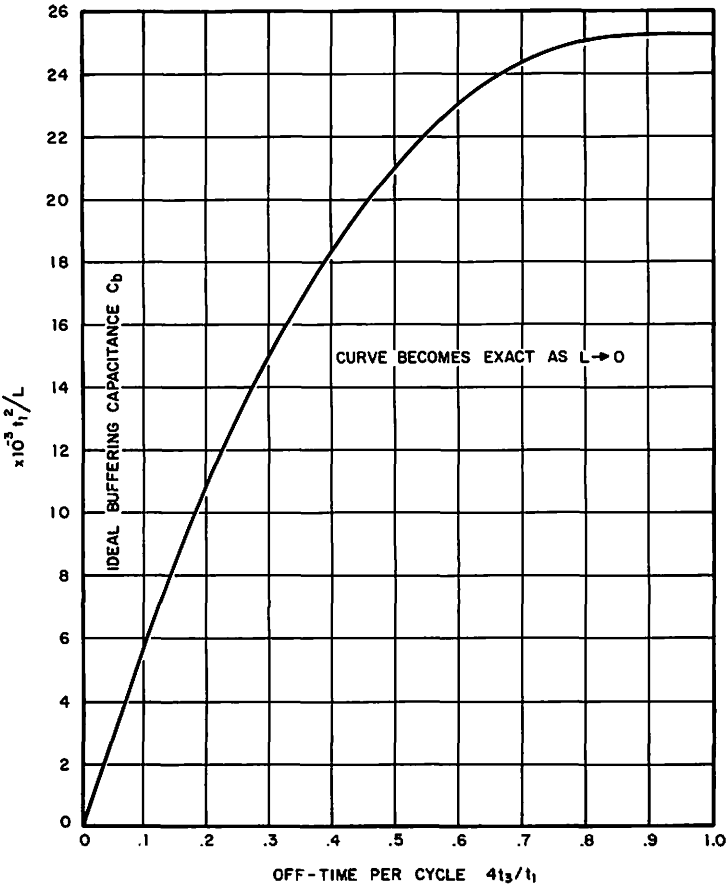


FIGURE 24. *Ideal buffering capacitance versus off time.*

Our theory so far sets no lower limit for L. But two practical factors do so, the first being winding resistance, and the second, errors in buffering.

As L is reduced, the magnetizing current increases and gives rise to greater primary losses, even if the primary resistance decreases with L. Suppose that L/R'_3 is several times less than $t_2/2$ so that we are well into the oscillatory region. Then the peak magnetizing current i_o, which equals $V_1t_2/2L$ is several times greater than V_1/R'_3, and the primary losses are

largely due to the magnetizing component of primary current. This is triangular, and the losses during one dwell period, taking $t = 0$ when $i = 0$, are:

$$2\int_0^{t_2/2} R_p\, i^2\, \delta t = 2\int_0^{t_2/2} R_p \left(\frac{i_0 t}{t_2/2}\right) \delta t$$

$$= \frac{1}{3}\, i_0^2 R_p\, t_2 = \frac{1}{12}\, V_1^2 R_p\, t_2^3/L^2$$

These losses become significant when they approach the energy supplied to R'_3 during the same period, namely $t_2 V_1^2/R'_3$, the gain being approximately halved when they are equal. Hence a lower limit for L is given by:

$$\frac{L}{R'_3} \cdot \frac{L}{R_p} \geqslant \frac{t_2^2}{12} = \frac{t_1^2}{48} \text{ approximately.}$$

This is still not very helpful if R_p is very low. However, the lower we make L, the larger is the ideal C_b and the greater the losses if buffering is not correct. Physically, this is because of the larger energy circulation involved. Some drift in contact adjustment is inevitable during the life of a chopper, and this upsets the buffering. Hence L/R'_3 must not be too low, and a minimum value in the region of $t_1/50$ is suggested.

If this condition is met, and R_p is made sufficiently low that $L/R_p > t_1$, then the waveform will have reasonably small droop, and primary losses will not unduly reduce the gain.

Again it is essential to test any particular design with the off time and the inductance varied through the full range of values which may be encountered, not only due to production spreads, but also due to variations over the chopper lifetime. Short off times are desirable, since they reduce C_b and the errors due to incorrect buffering.

DESIGN EXAMPLE. Taking the same initial requirements as we did in the MBB example, we have:

$t_1 = 20$ m.s.
$t_2 = 9$ m.s.
$t_3 = 0.5$ m.s.
$R_1 = 1000$ ohms.
$R_3 = 100{,}000$ ohms.
k (opt) $= 10$.
Gain (opt) $= 10$.
$R'_3 = 1000$ ohms.

and the suggested range of L becomes 2 to 0.4 henrys for oscillatory conditions. This is less inductance than was required in the MBB case, and the winding resistance should be negligible in practice, so that the gain will in fact, approach 10.

SECTION VI

SOME INPUT CIRCUITS

A frequently used modulation method is the circuit of Figure 25, a half wave modulator. The input is periodically shorted to ground, and the resultant pulse is shown. The waveform shown represents negative DC input (because the pulse is positive going), and the peak-to-peak wave height at the input grid is about the DC value at the amplifier input. Often the remaining contact is used to demodulate at the amplifier output.

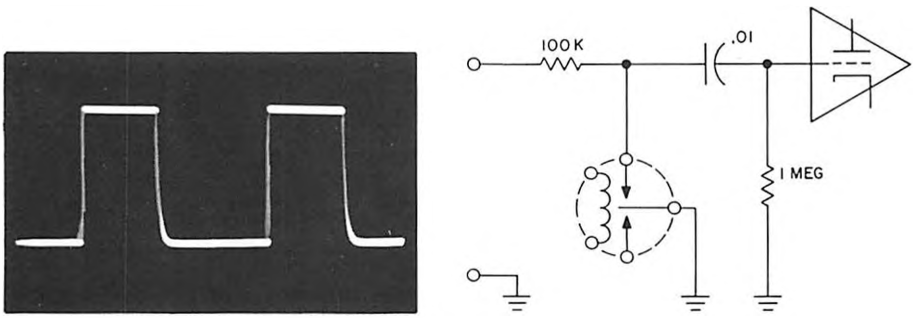


FIGURE 25. *Waveform of a half wave modulator.*

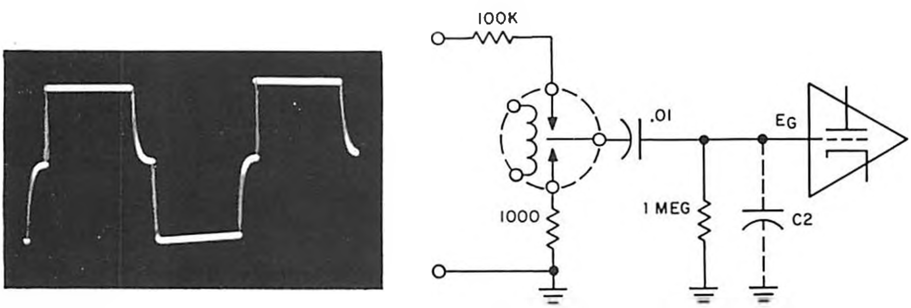


FIGURE 26. *Full wave modulator showing capacitive effect during off time.*

A slightly more sophisticated circuit, but one which ties up the other contact, is shown in Figure 26. (The other contact might be used for a rectifier, or to modulate another amplifier.) During the transit time of a break-before-make chopper in the circuit of Figure 26, the

amplifier looks into a high impedance, the effect thereof being noticeable in the off-time curvature (discharge of C_2, stray circuit capacity). Therefore, there is some danger of noise pickup during off time, if attention is not paid to noise reduction. The output voltage E_g of Figure 26 will have a peak-to-peak value about twice that of Figure 25, assuming circuit components are the same in both cases.

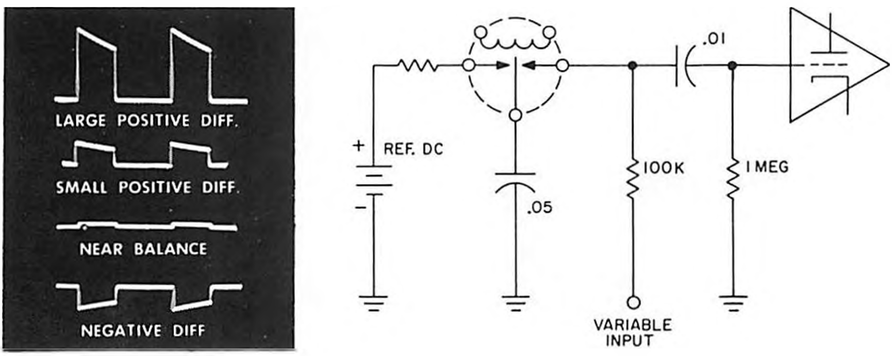


FIGURE 27. *Comparison circuit examines difference of two DC values.*

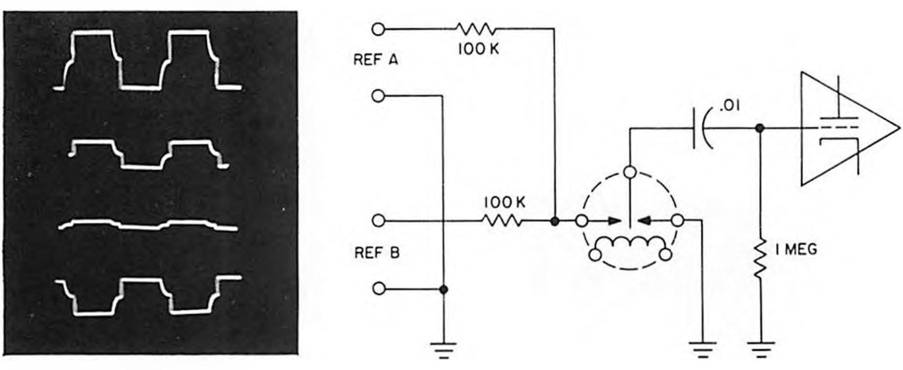


FIGURE 28. *Summing circuit adds signals A and B.*

A less satisfactory solution is the use of an MBB chopper, which maintains contact continuity. Of course, this makes us dependent on perfect contact action. Perfection is not always obtainable, and while make-before-break choppers are readily available, it may be better to design for maximum reliability. If the circuit is unaffected by an open contact, either kind of chopper can be used at will.

It should also be noted that with an MBB chopper, during the common time, all contacts are common and the input is grounded. If

this causes undue loading on the input, it may be necessary to insert resistance in series with the grounded contact, which will in turn reduce the circuit gain.

COMPARISON. A frequent application requires the comparison of two DC values, such as a reference standard of voltage and an unknown so that the difference signal will be zero if the two values are equal. Figure 27, is a typical circuit with the amplifier input waveform shown.

The charge on the capacitor from the standard source is compared at the grid of the input stage with the unknown DC signal to be measured, and the amplifier observes the difference between them. This is an excellent zero hunting circuit and the basis for some successful servo systems.[1,2]

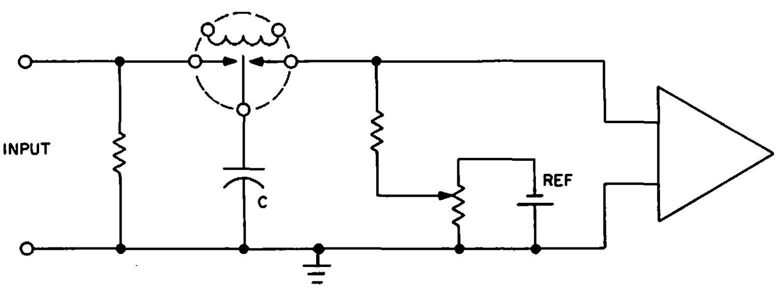


FIGURE 29. *Series summing circuit.*

Signals need not be compared in two contacts. They can be mixed by the use of isolating resistors and examined as in Figure 28; as inputs A and B are varied, the resultant patterns appear as photographed.

Figure 29, is a variation of Figure 28, in which a sample of the unknown is voltage summed and (usually) servo driven to zero. The chopper periodically samples the signal on the input side and delivers it to the amplifier and reference in series. All of these circuits, with many variations, will be found in digital voltmeters.

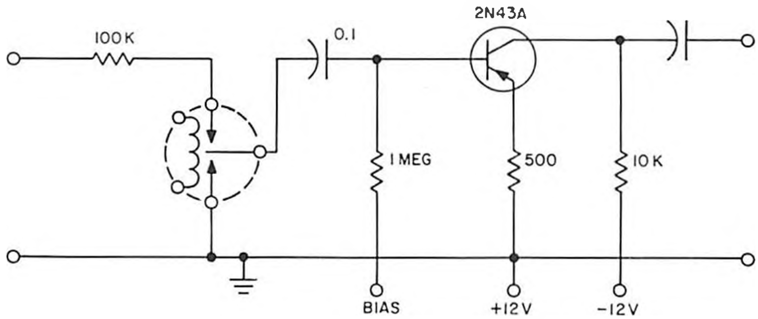
Still another comparison circuit is shown in Figure 31, resulting in a totally different pattern. The wave halves are substantially independent of each other, as will be observed from the waveforms. In the upper pattern one is positive, one negative; in the lower, both inputs are negative.

TIME SHARING. Figure 31, gives the possibility of using one amplifier to perform two functions. If synchronous demodulation (as for example, another chopper) is employed to restore the information ob-

[1]Digital Presentation Vacuum Tube Voltmeter. Nuut and Munsey, Electronics, January 1956.

[2]Design of a Ratiometer. Kuehn, Electronic Equipment, November 1955.

tained at inputs A and B. it is possible to perform two essentially independent operations. It should be noted that, since an AC amplifier is used, a new zero or base line is established, and without special clamping a demodulator may accept part of both signals. This arrangement has been used deliberately when the two signals were interdependent.



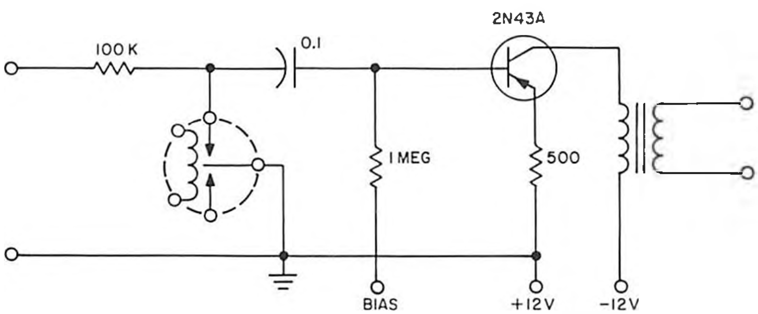


FIGURE 30. *Modulation circuits for transistors are substantially the same.*

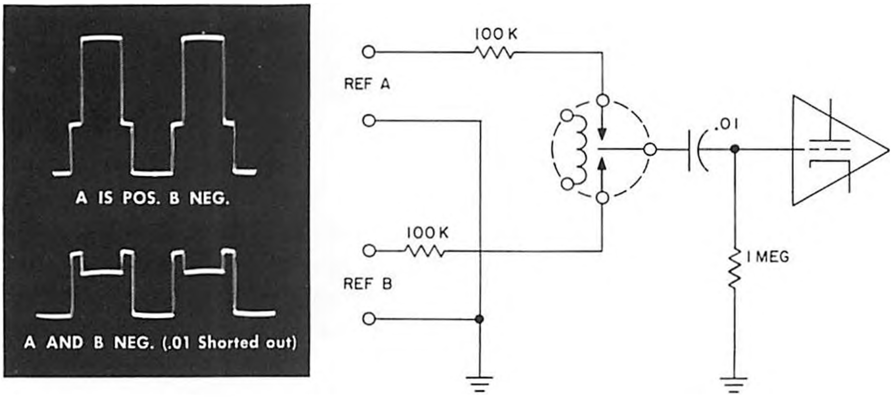


FIGURE 31. *Comparison or time sharing circuit.*

HIGH INPUT VOLTAGES. Comparison circuits sometimes are called upon to examine rather substantial voltages. Some chopper manufacturers limit certain of their models to a low voltage, probably due to contact welding or sticking if the drop across an open contact becomes excessive. Other manufacturers, AIRPAX included, permit voltages as high as 100 volts or more, for some models.

The difference[3] is one of structure. If the contacts are very small and the physical motion very minute, obviously the contacts will be easily damaged. There are other problems with such structures, such as

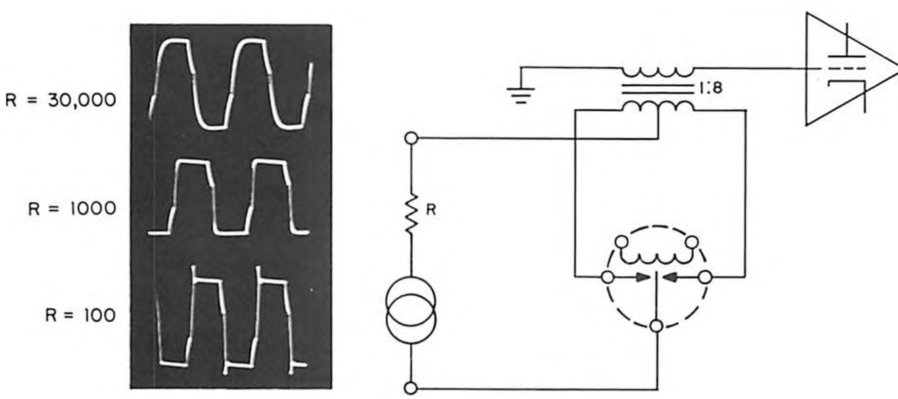


FIGURE 32. *Balanced input using a transformer; waveforms are for 3 values of source resistance.*

reduced resistance to extremes of temperature. On the other hand, minute motion and light contacts permit a high self-resonant frequency and operation over a wide frequency range.

INPUT ISOLATION. The transformer input circuit of Figure 32, has many virtues as was observed[4,5] some years ago, including the ease of obtaining a floating input (both input terminals free of ground connection). It is in fact almost the only possible way, as other methods require some form of ground return. The transformer, besides providing isolation, permits considerable voltage gain. The observed waveform will contain quite a few reactive effects, as illustrated. The excessive curvature in the upper pattern is due to distributed capacity reflected into the

[3]Considerations of Relay Dynamics With An Example of Non-Linear Vibrating Reed Design. David A. Robinson, presented at Fifth National Conference on Electro-Magnetic Relays, Oklahoma Institute of Technology, Oklahoma A & M College, Stillwater, Oklahoma.

[4]A Contact Modulated Amplifier to Replace Sensitive Suspension Galvanometers. M. Liston, et al, Review of Scientific Instruments, Vol. 17, No. 5, May 1946.

[5]D-C Amplifier Stabilized for Zero and Gain. Williams, Jr., Tarpley and Clark, AIEE Transactions, 1948, Vol. 67.

primary. (The transformer, designed for choppers, had a step-up ratio of 1:8. See Figure 33.) In the lowest picture of Figure 32, there appears some overshoot or "ring" at the leading edge of the wave. Reproduction of the square wave can be improved considerably by the use of AC feedback around the AC amplifier portion only; in fact, this offers the opportunity of achieving a high order of linearity without DC feedback and consequent loss of the input "float."

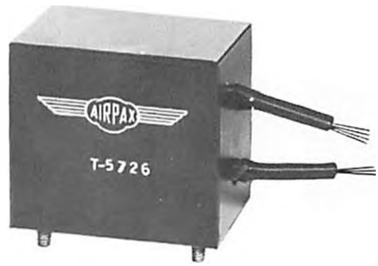


FIGURE 33. *Transformer designed for chopper use.*

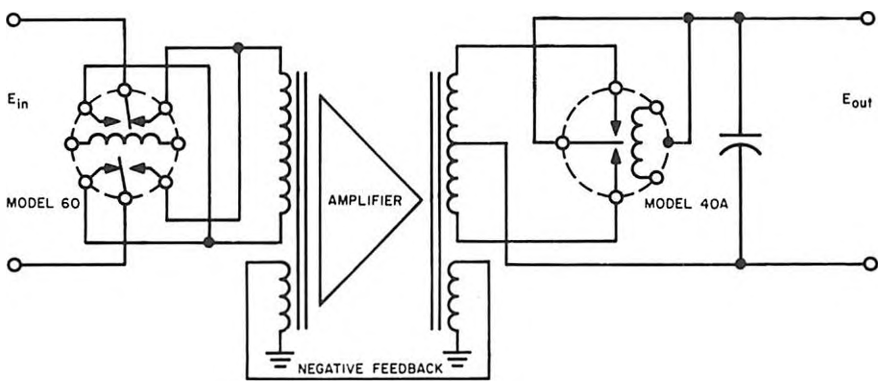


FIGURE 34. *Techniques to reach one microvolt signals.*

This circuit also permits, by careful attention to input balance, a high order of rejection to a signal between both input terminals and ground, reputedly reaching as high as a million to one. This is the "common mode rejection" given frequently as a virtue for data logging amplifiers, and is discussed in later Parts. A typical application is a long line run from a thermocouple delivering only a few millivolts. Strong AC fields may produce signals of many times that figure, but from both lines to ground. If impedances to ground are balanced, the noise signal cancels.

There are some very distinct advantages in the use of full wave

modulation, transformer input, with a DPDT type of chopper, as shown in Figure 34. This is most particularly true at the lowest signal levels. A first advantage is that the transformer primary has four times the impedance of the more usual center tapped modulator circuit. The specifications always seem to demand the highest possible impedance.

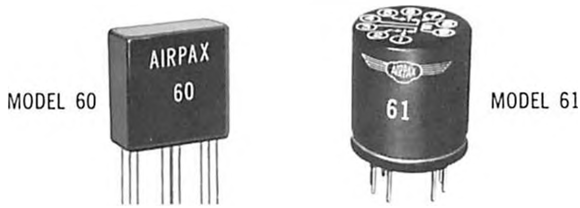


FIGURE 35. *Miniature double pole double throw choppers.*

Low noise levels are also easier to obtain with a DPDT chopper unless the chopper itself is noisy. It is rather simple to arrange the transformer primary in two identical halves, so that both primary leads are perfectly symmetrical to ground. It is considerably more difficult to provide this precise a balance when the center tap circuit is used. Chopper Models 60 and 61, as shown in Figure 35, being also perfectly symmetrical as well as low noise, continue the perfect balance out to the input terminals of the amplifier. The net effect is remarkably good common mode rejection. Since the noise in these choppers is virtually nonexistent, the recognition and use of signal levels of a few microvolts becomes feasible, even in strong noise fields.

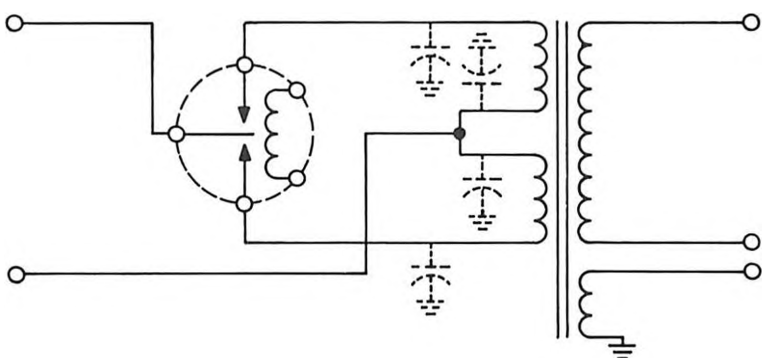
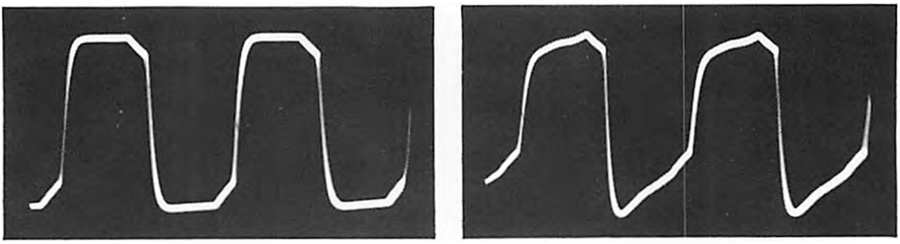


FIGURE 36. *Capacitive balance is difficult with a center tap modulator.*

Figure 36, will illustrate readily that with the center tap circuit the capacitive balance at the input is likely to be less than perfect, reducing

common mode rejection. In addition, as described in a later section, the thermal differential (offset) problem is less likely, since the circuit is perfectly symmetrical.

The "initial" permeability of the input transformer core presents a problem, as shown in Figure 37. The permeability of some materials falls



HIGH LEVEL LOW LEVEL

FIGURE 37. *Poor initial permeability may limit transformer performance.*

off seriously as the level approaches zero. Fortunately the published curves of 80-20 nickel-iron show good permeability even at 0.1 gauss, and one assumes hopefully it will still be good at 0.001 gauss. At impedances of 1,000 ohms the power level, E^2/R, is 10^{-15} watts, and the use of the high permeability alloys is mandatory. It helps to use more turns, too, since impedance varies as the square.

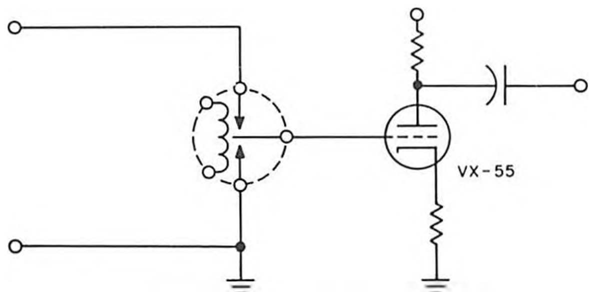


FIGURE 38. *Direct coupling has some value at very low grid current in high impedance circuits.*

HIGH IMPEDANCES. A good quality capacitor is required at the input grid of most chopper amplifiers. Any leakage will be chopped and result in noise or offset, i.e., unwanted signal. There are times, however, when very high impedance circuits are desired and the signal levels are also rather high. Electrometer tubes, such as type VX-55, have exceptionally low grid current and can be direct-coupled in the manner of Figure 38 for special applications, for example, as a detector for an in-

sulation resistance bridge. Grid current is no doubt present in the VX-55, however in this application it was sufficiently low to avoid being troublesome.

TRANSISTOR CHOPPERS. The appeal of solid state devices is almost universal among engineers and the idea of a solid state chopper is certain of a welcome. "Solid State" is often so desired, perhaps as a matter of pride, that it is likely to be used even in spite of poorer performance. A frequent example is the high input impedance transistor circuit; almost any tube has a higher impedance input, at lower cost and with much greater dynamic range.

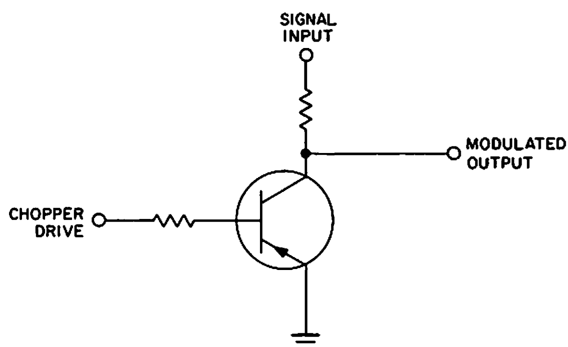
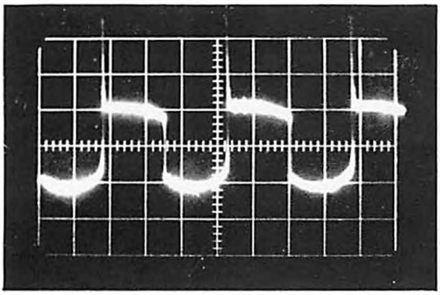


FIGURE 39. *Basic transistor chopper circuit.*

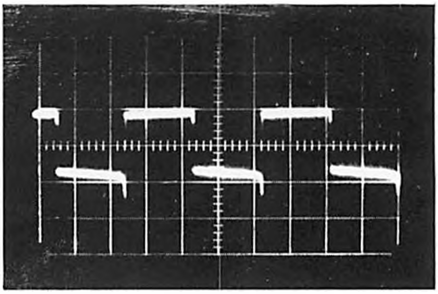
However, there are practical uses for a transistor chopper. It will permit offset and drift about twice that of the best DC amplifier, perhaps a little more. It will also stabilize a pretty bad DC amplifier. It cannot be sine wave driven for best performance, as will be apparent from noise illustrations later shown. If the drive is locally generated, this is no hardship, square waves being difficult to avoid. A transistor chopper is generally assumed to be more reliable. This is likely to be true; it is at least as reliable as the rest of the semiconductors. Unfortunately, when the associated parts such as a driver transformer are considered, it will be bigger, heavier, and more expensive than the best of mechanical choppers. A good engineer will weigh expense and performance and will use that best fitting his problem. If the sales department demands solid state, try to use a transistor chopper. If they demand lowest cost, use mechanical. If they ask for the limit of the state of the art, also use a mechanical chopper. This will not be the same chopper you chose for lowest cost.

BASIC DESIGN AND OPERATION. In its most basic form, a transistor chopper would appear as shown in Figure 39. The transistor is alternately biased on (to saturation) and off by a square wave drive. This permits current flow from the signal source to the load to be interrupted at the frequency of the drive.

With this simple arrangement, several undesirable characteristics would be obtained. When the transistor is biased on, even fully saturated, a small potential difference exists between its collector and emitter, ranging from about one to one-hundred millivolts for different transistors.



(a) 140 mv peak-to-peak noise output with sine wave drive.



(b) 40 mv peak-to-peak noise output with square wave drive.

FIGURE 40. *Transistor chopper noise output with zero input signal. Load 10,000 ohms.*

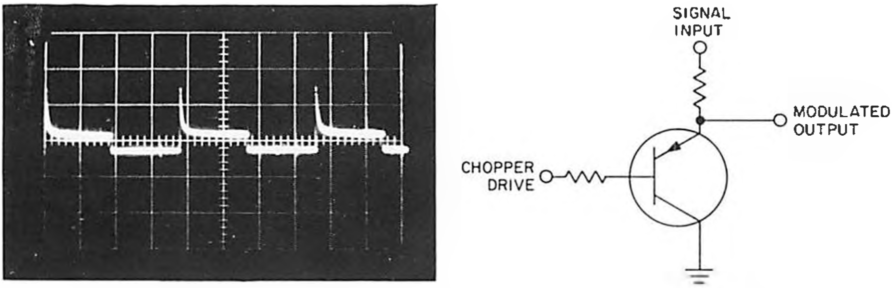


FIGURE 41. *Inverted connection has lower noise output.*

Figure 40, illustrates an extreme case, using a 2N398, where the noise output at zero signal is about 140 mv peak to peak. This output will, of course, become offset upon demodulation. Considerable improvement is immediately possible using "chopper" transistors, by the inverted connection of Figure 41, and finally by the use of a clean square wave drive. Figure 41, pictures noise of about 1 mv rms, using 6 volt square wave drive, a 10K output load, under zero signal conditions. A more effective remedy for this substantial offset voltage may take the form shown in either Figure 42 or Figure 43. In Figure 42, the transistors are both on at the same time, with their saturation voltage drops in opposition. If the transistors are matched for equal voltage drops, there is nearly complete cancellation. If the circuit is re-arranged in the inverted circuit, Figure 44, still further improvement can be had.

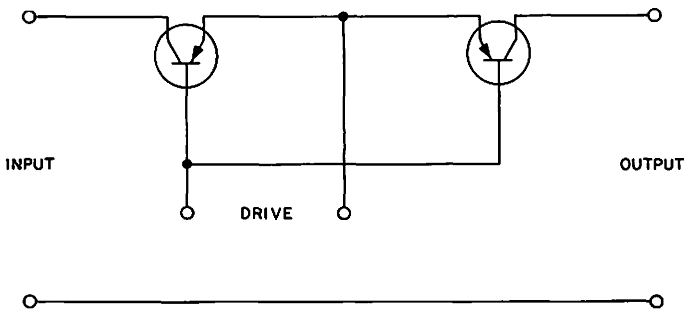


FIGURE 42. *Bilateral transistor chopper, SPST.*

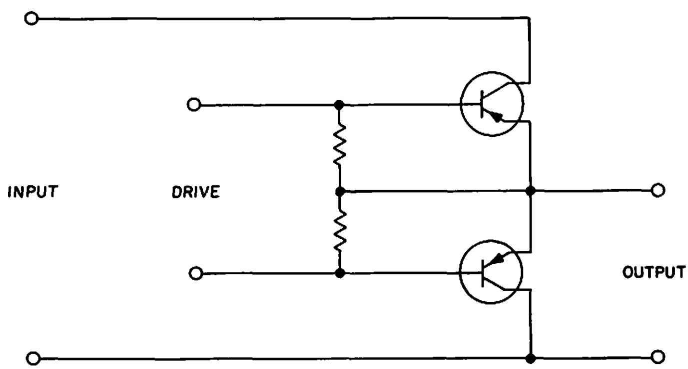


FIGURE 43. *Series-shunt transistor chopper, SPST.*

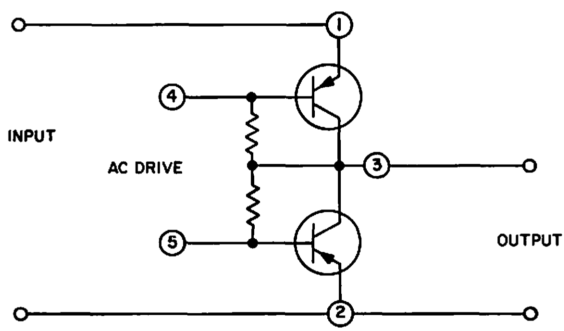
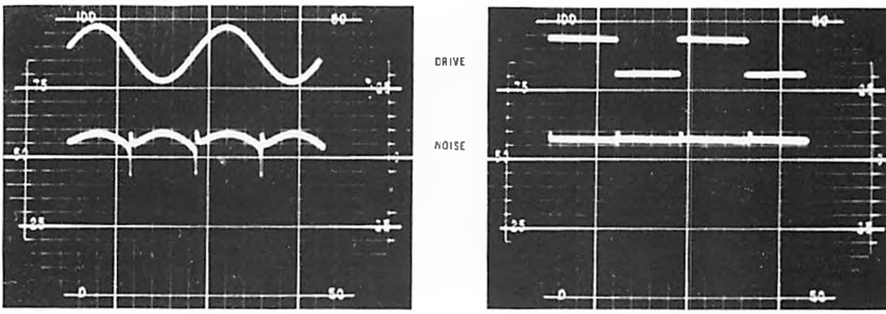


FIGURE 44. *Series-shunt arrangement, inverted connection.*

In Figures 43 and 44, the transistors are back to back and driven out of phase so when one is on, the other is off. The result is that alternately the voltage drop of one or the other transistor is coupled to the load with each transistor on for nearly one-half cycle. If the transistors are perfectly matched for equal drops, the result is a constant DC coupled



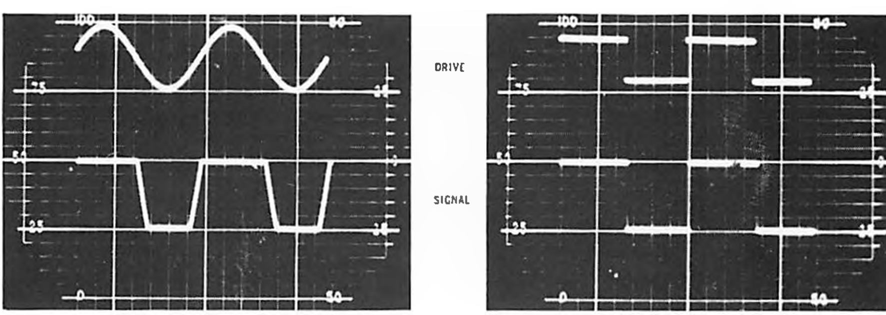
(a)

Noise voltage. Circuit: Figure 44. Scale: 2 mv. Load: 10,000 ohms. Sine wave drive.

(b)

Noise voltage. Circuit: Figure 44. Scale: 2 mv. Load: 10,000 ohms. Square wave drive.

FIGURE 45.



(a)

Signal voltage. Circuit: Figure 44. Scale: 5 volts. Load: 10,000 ohms. Signal chopped: 5 volts DC.

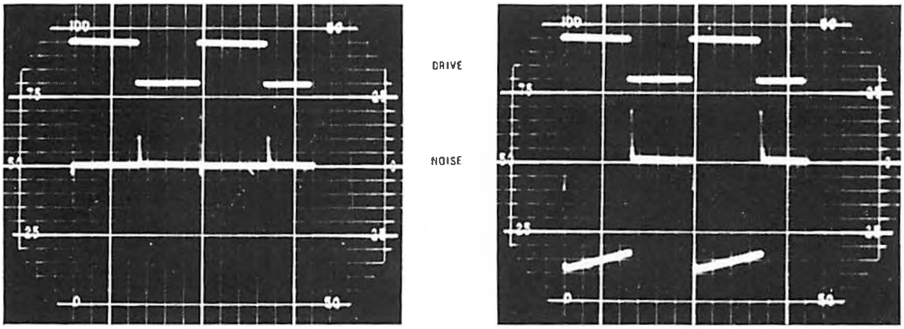
(b)

Signal voltage. Circuit: Figure 44. Scale: 5 volts. Load: 10,000 ohms.

FIGURE 46.

to the load with no offset component. The results of this approach are shown in Figure 45. In both cases, the DC component is evident. Here the transistors were both connected inverted, but the normal connection would give similar results except that both the ripple and DC component would be greater. Note that both offset and ripple appear substantially higher in the case of the sine wave drive than with a square

wave, as is of course to be expected with transistor choppers. The wave shape of the modulated signal voltage is shown in Figure 46. (In both Figures 45 and 46, the drive signal appears as the upper wave, the chop-



(a)
Noise voltage. Circuit: Figure 42. Scale: 2 mv. Load: 100 ohms.

(b)
Noise voltage. Circuit: Figure 42. Scale: 2 mv. Load: 10,000 ohms.

FIGURE 47.

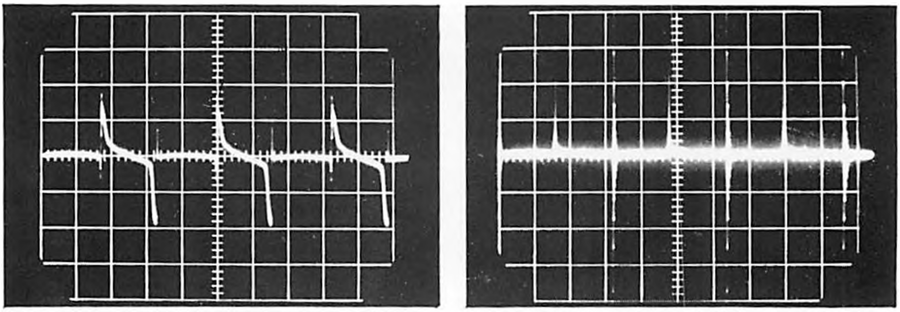
per output as the lower pattern.) The distortion and noise resulting from a sine wave drive, make a square wave drive a necessity unless limited performance is acceptable.

Another characteristic of the chopper in Figure 39, which may be troublesome, is the result of the collector cutoff current of the transistor. This small current will pass through the load and, where the load impedance is high, can produce a substantial offset voltage. This problem is reduced by using the circuits shown in Figures 42 or 43, if the transistors are *also* matched for equal cutoff currents. In Figure 42, the currents occur at the same time but in opposite directions and tend to cancel. In Figure 43, the transistors take turns in supplying this current to the load, also cancelling any offset which might be caused.

Two more factors may be helpful. Since the cutoff current has a high source impedance and is, therefore, independent of impedances external to the transistor, the voltage drop produced may be made small by making the load impedance as small as practical. Also, at any given temperature, the cutoff current of a silicon transistor will be much less than its germanium counterpart. The effect of load impedance in the circuit of Figure 42 is demonstrated in Figure 47.

It is also necessary that matching of pairs be carefully accomplished, for maximum performance, and it may prove necessary to match pairs over a range of temperatures. The use of integrated circuit techniques and micro-transistors closely paired on a common wafer of silicon may offer considerable improvement.

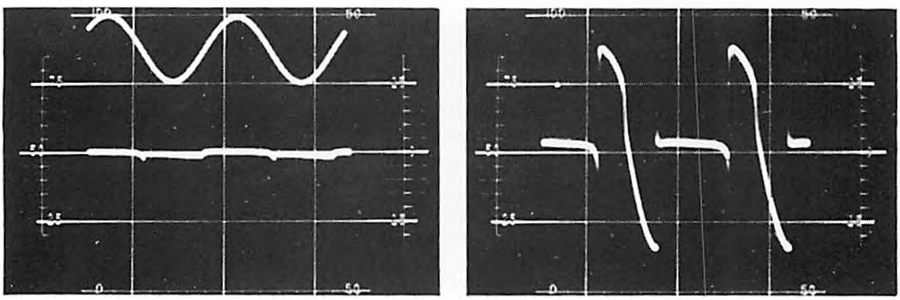
The linearity using an arrangement of the form in Figure 42 is limited at the low level end by offset and noise. Offset can be adjusted out elsewhere in the system, and in most cases, signal to noise ratios as low as unity still give good results since the noise and the resultant offset are constant. One more characteristic which should be considered is the effect of junction capacitance. This capacitance provides an undesired



(a) *Sine wave drive. Circuit: Figure 44. Scale: 2 mv/cm.*

(b) *Square wave drive. Circuit: Figure 44. Scale: 2 mv/cm.*

FIGURE 48.



(a) *Noise voltage. Circuit: Figure 42. Scale: 2 mv. Load: 100 ohms.*

(b) *Noise voltage. Circuit: Figure 42. Scale: 20 mv. Load: 10,000 ohms.*

FIGURE 49.

means of coupling drive voltage into the load. Since the capacitance is voltage dependent and is maximum at the instant the transistor is being turned off, a spike of considerable amplitude may result, Figure 48. This spike can be suppressed by limiting the amplifier bandwidth (filtering). There is also a quadrature component of the drive voltage which reaches the load. This produces no offset in itself but makes a noise component which may have a masking effect on small signals. With such a small

capacitance (i.e., high reactance) the noise produced will be proportional to the load impedance and to the drive frequency used. Figure 49, shows the effect of load impedance. (Note the difference in scales.) Where small signals must be considered, this effect becomes a limitation at much lower frequencies than those related to actual switching time. Thus, the upper chopping frequency limit for low level use may not exceed 5,000 cycles.

While these characteristics may seem to restrict the usefulness of transistor choppers, their advantages may exceed their limitations. Since they use stationary parts, they are not disturbed by extremely high shock, vibration or acceleration. They perform with very small lag so they

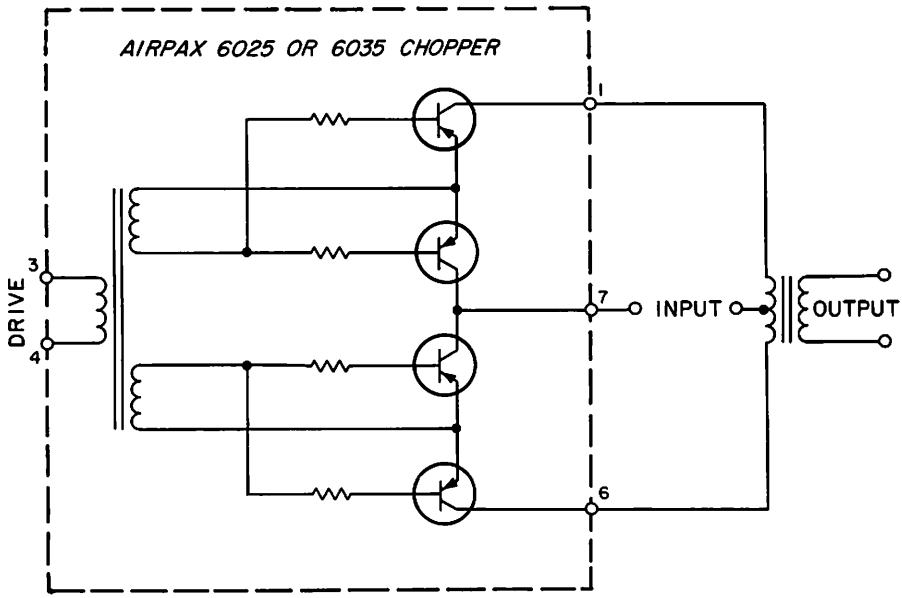


FIGURE 50. *Bilateral SPDT transistor chopper used as a full wave modulator.*

operate in phase with their drive voltage, avoiding external correcting networks. Both phase angle and dwell time are nearly independent of drive frequency, voltage, and temperature. And they do help to provide the magic title of "all solid state."

Figure 50 shows a typical application using the techniques of Figure 42. Two pairs of bilateral switching transistors are combined to form a SPDT chopper used in this case in a balanced transformer input scheme. The SPDT choppers with matched transistors and complete with isolating drive transformer are commercially available in a miniature 7-pin base can (AIRPAX Types 6025 and 6035). Modulation and demodulation of

input and output for an operational amplifier are shown in Figure 51. Figure 52 illustrates a variety of transistor choppers based on the preceding concepts. Chopper Types 6010 and 6011 will require a separate external transformer. The performance of Type 6010 can be observed from the circuits and patterns illustrated in Figures 53 and 54. The deterioration at high frequencies is evident.

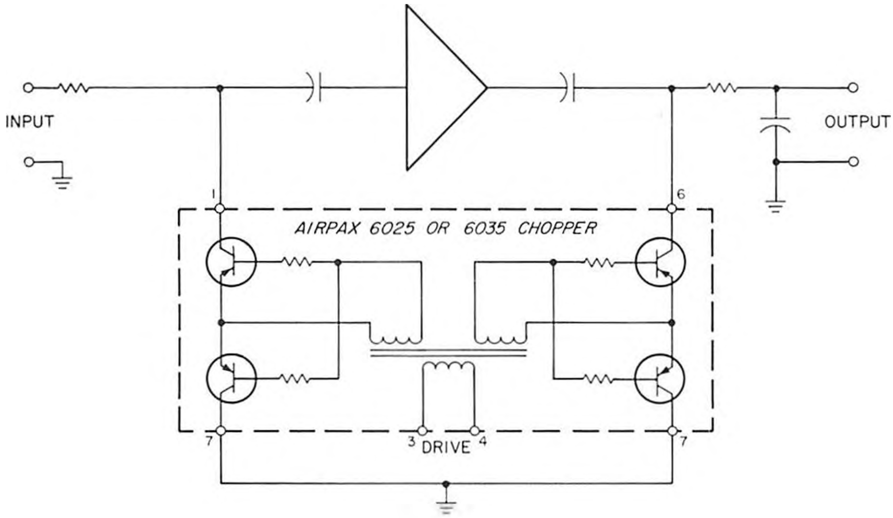


FIGURE 51. *Bilateral SPDT transistor chopper used as an input modulator-output demodulator.*

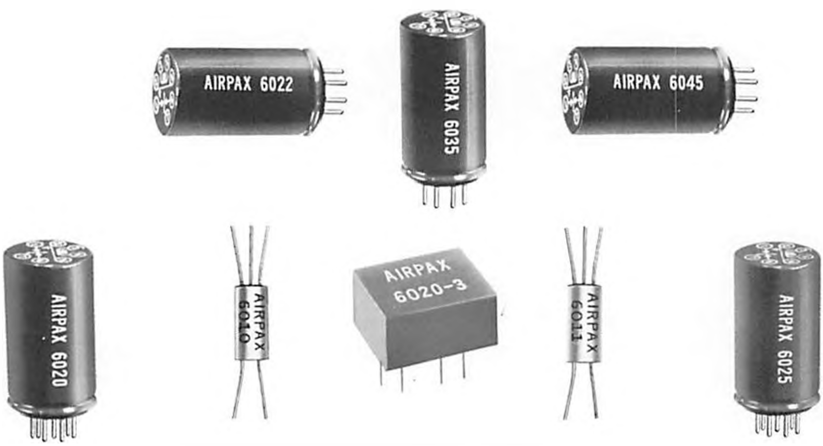
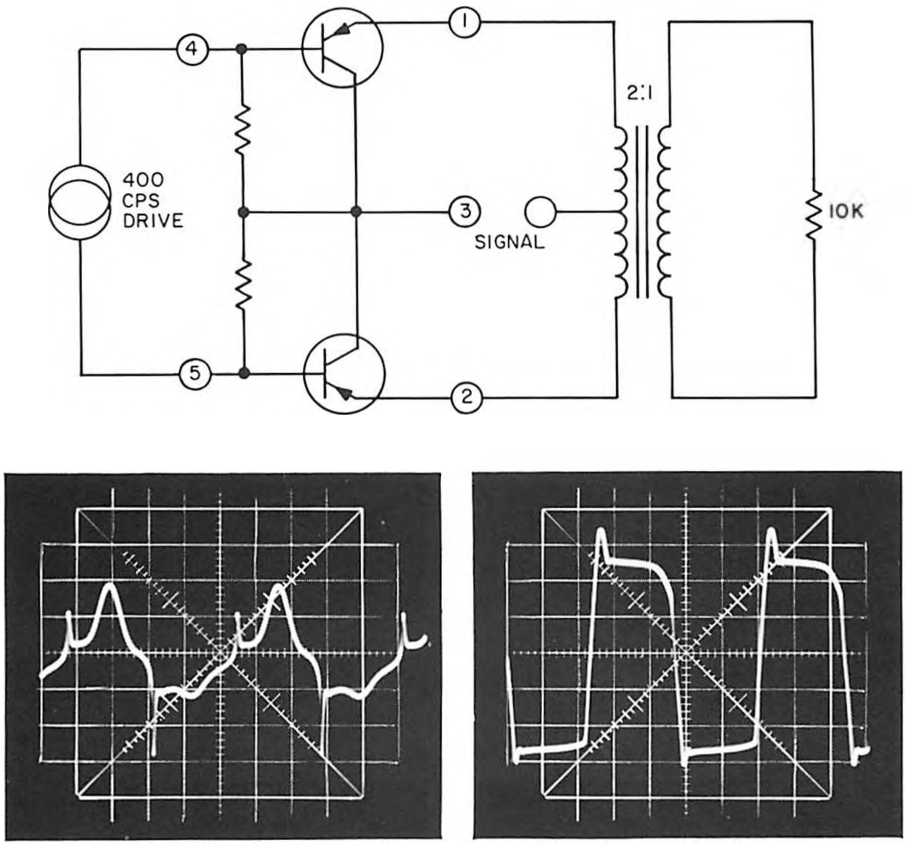


FIGURE 52. *Transistor choppers using the series-shunt and bilateral circuits.*

The foregoing is a basic description of the design and operation of the standard transistor chopper. The approach has been to match a pair of transistors and drive them with a subminiature audio transformer and resistors, supplied as either a separate unit or as a part of the chopper package.

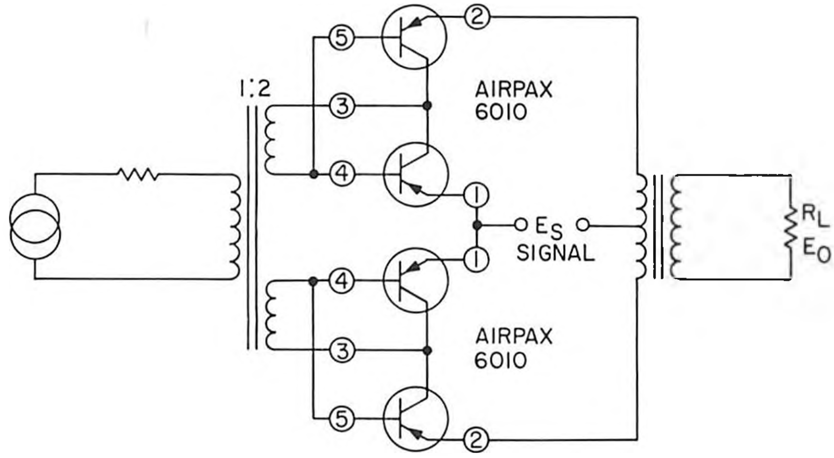


1 mv/div. vert., 0.5 msec/div. hor.
(a) Output at null

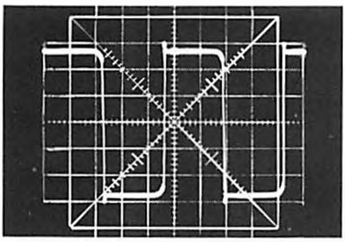
0.5 v/div. vert., 0.5 msec/div. hor.
(b) Output with signal

FIGURE 53. *Modulator circuit employing Type 6010 transistor chopper.*

There is an additional technique (AIRPAX patents pending), in which an RF carrier is used to provide the isolation of drive and contacts imperfectly supplied by the driver transformer used in the preceding circuits, and this is the basis of AIRPAX Series 7001 choppers. This is one of the difficulties with transistor choppers, i.e., the intimate and necessary connection of drive to signal. In mechanical choppers, this isolation can be provided almost completely by mechanical means.

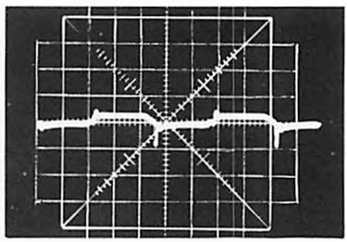


Full wave modulator improves the null stability.



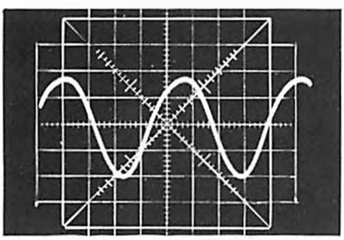
$R_L = 10$ K $E_S = 1.5$ v
0.5 v/div. vert., 0.5 msec/div. hor.

Signal voltage waveform.



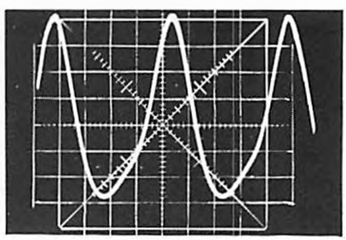
$R_L = 10$ K $E_S = 0$
1 mv/div. vert., 0.5 msec/div. hor.

Noise value of 0.19 millivolts.



$R_L = 10$ K $E_S = 100$ mv
50 mv/div. vert., 2 μ sec/div. hor.

(a)



$R_L = 10$ K $E_S = 0$
10 mv/div. vert., 2 μ sec/div. hor

(b)

With 100 KC drive (a) E_o output of 46 millivolts, (b) null value of 15.7 millivolts.

FIGURE 54.

As shown in Figure 55, the drive, preferably square wave for the same old reason, is applied to terminals 1 and 2, providing the energy for a Colpitts RF oscillator operating at approximately 2 MC, the energy from which is then demodulated by diode D1 and switches transistors Q_2, Q_3.

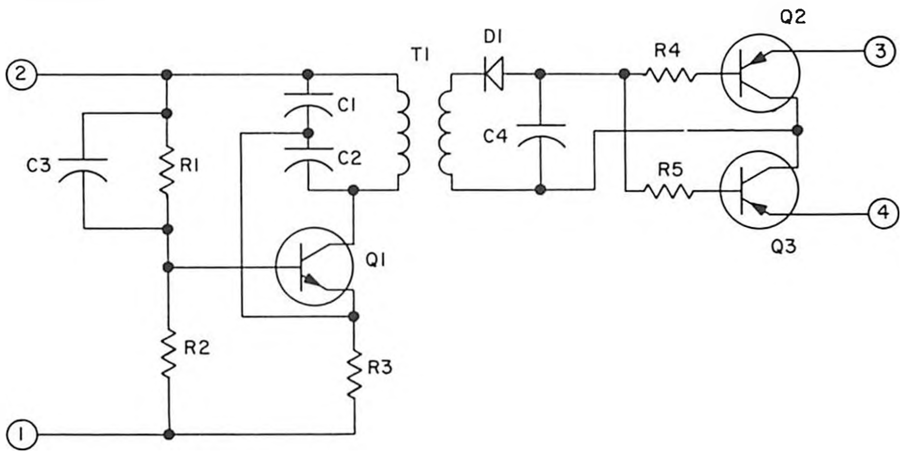


FIGURE 55. *Schematic of chopper Type 7001.*

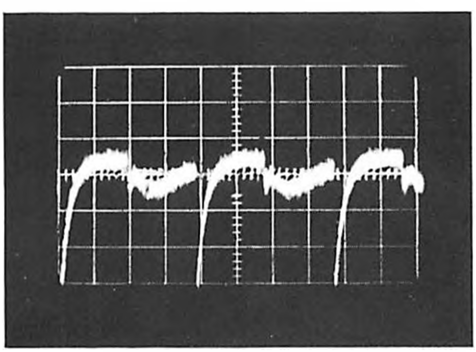


FIGURE 56. *7001 Series transistor chopper. Noise across 10 K. Drive: 400 cycle square wave. Scale: 30 μv/cm rms.*

This technique provides improvement in performance of one order of magnitude, perhaps a little more, and residual noise levels as low as 35 microvolts rms across a 10,000 ohm load become possible. Figure 56 is the noise observed at the output of a typical Type 7001 half wave chopper. The spike noise reaches about 100 microvolts peak to peak, but the effective noise is about 30 microvolts, and this will be the offset contribution of the chopper.

Input-output linearity is in the vicinity of ½%. Transistor choppers in general may exhibit nonlinearity, at low circuit impedances in particular. The maximum signal is ±7 volts peak, a substantial value permitting a wide dynamic range. Temperature drift is about 2 microvolts per degree C between −30°C and +100°C. This is a function of matching and will improve as the state of the semiconductor art improves.

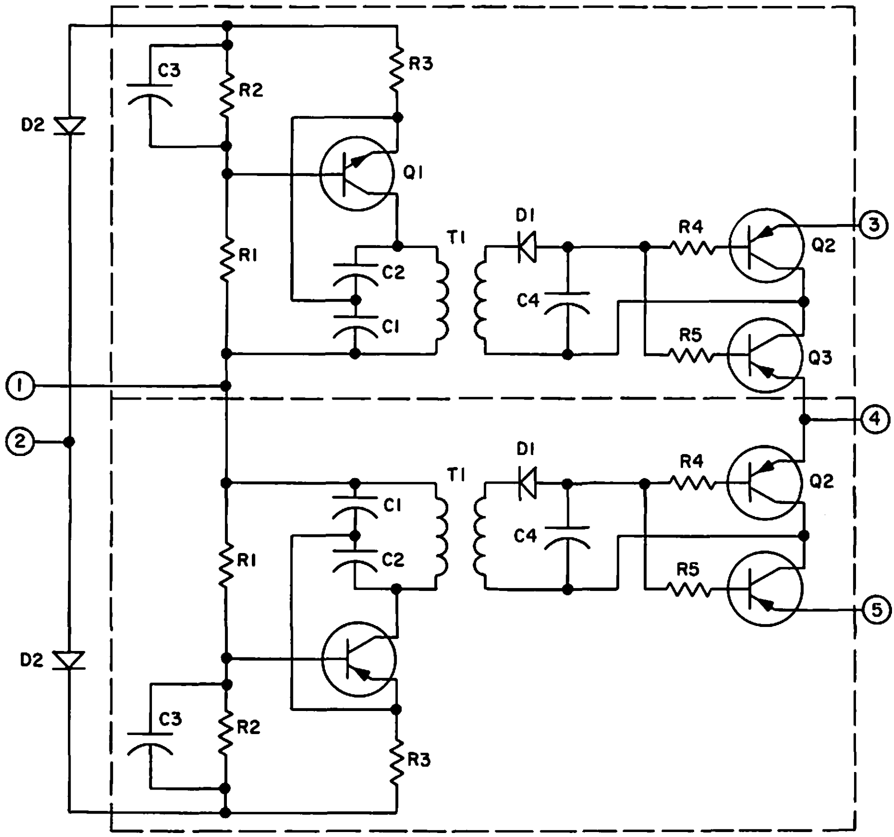


FIGURE 57. *Type 7005 SPDT chopper.*

The SPDT configuration is supplied by Type 7005, essentially two Type 7001 choppers, Figure 57. Figure 58, shows the modular assembly of these choppers, epoxy molded for printed circuit assembly.

The table following gives performance limits for a Type 7001 SPST chopper, and Figure 59 describes typical performance of a Type 7005, SPDT, under exposure to a variety of environmental conditions.

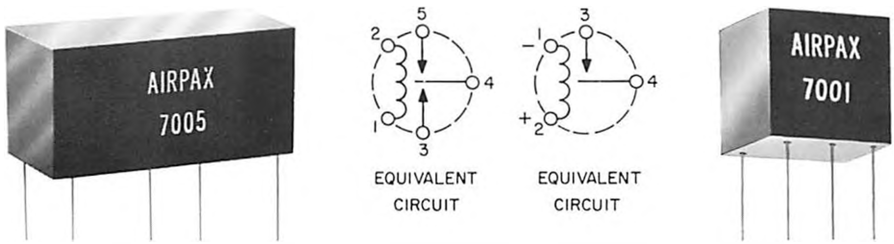


FIGURE 58. *High performance transistor choppers.*

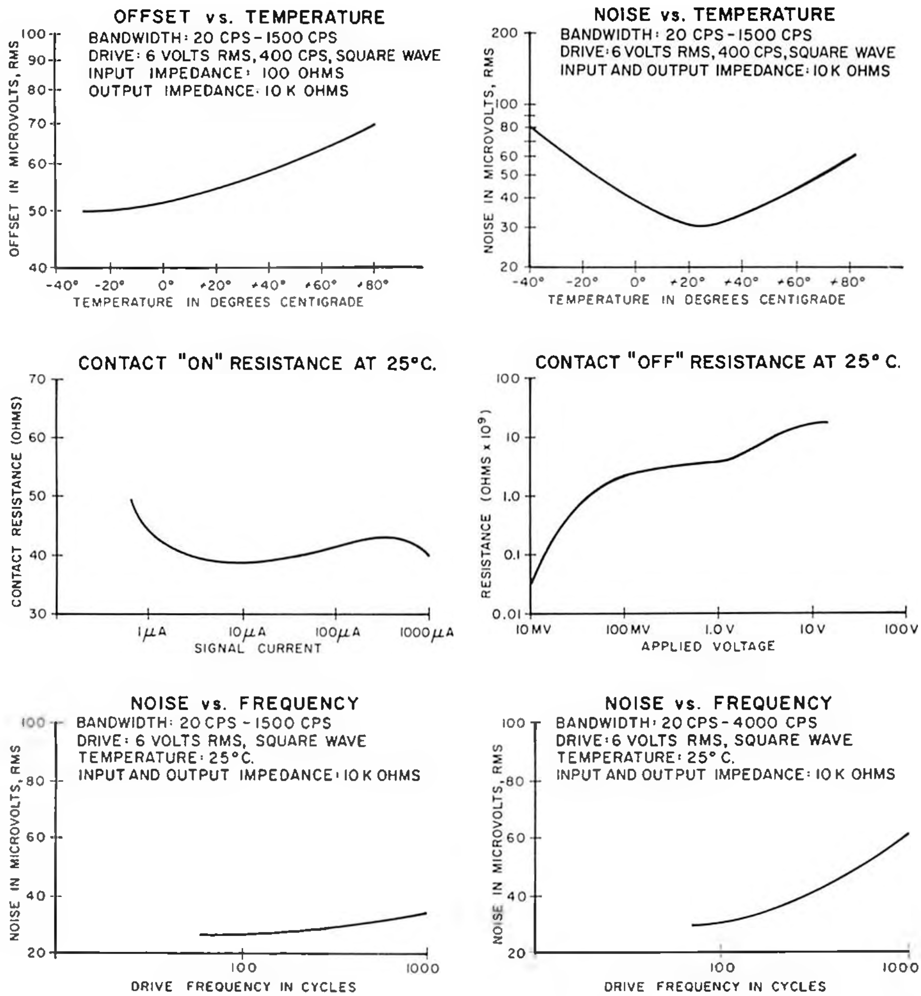


FIGURE 59. *Typical curves, Type 7005 SPDT transistor chopper.*

SPECIFICATIONS
(at 25°C unless noted)

PARAMETER	CONDITIONS	TYPICAL	MAX
Turn on Time	0 to 1 KC 1.5 V DC input sig. 10 K load 50 KC bandwidth	10 usec.	20 usec.
Turn off Time	0 to 1 KC 1.5 V DC input sig. 10 K load 50 KC bandwidth	20 usec.	30 usec.
Linearity	6 V DC drive 10 K load 50 KC bandwidth	0.5%	1.5%
On Resistance	10 K load 10 mv to 15 V DC input sig. 6 V DC drive	50 ohms	
On Resistance at −30°C	10 K load 10 mv to 15 V DC input sig. 6 V DC drive	50 ohms	
Off Resistance	No drive voltage 10 K load 0.1 V DC input sig.	1,000 meg.	
Off Resistance at +85°C	No drive voltage 10 K load 0.1 V DC input sig.	25 meg.	
Drive Input Impedance	6 V DC input to drive	1,200 ohms	
Noise	6 V 400 cps sq. wv. drive 10 K input and output load 20 cps to 1.5 KC bandwidth	35 uv rms	70 uv rms
Offset*	6 V 400 cps sq. wv. drive 100 ohms input 10 K ohms output 20 cps to 1.5 KC bandwidth	35 uv rms	70 uv rms

*That portion of the noise which is in phase with the square wave drive voltage.

TABLE 6. *Type 7001 performance limits.*

OTHER MODULATORS.[6] A complete description of the many modulation methods is beyond the scope of this text. Almost innumerable variations occur to the fertile minds of engineers involved in DC systems.

Magnetic modulators offer some promise, being quite stable and possessing a magnificent reputation for high reliability. However, if one is sufficiently concerned with reliability to need the probable 100,000

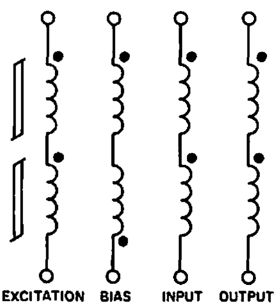


FIGURE 60. *"Fundamental" type magnetic modulator. Response is slower as sensitivity is increased.*

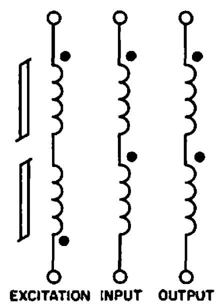


FIGURE 61. *"Second Harmonic" type magnetic modulator. The frequency response is limited.*

hours[7] of life obtainable from magnetic devices, it seems sensible to use a magnetic amplifier rather than transistor or tube types with magnetic modulation.

The mechanical chopper out-performs tubes and probably transistors, with regard to life and general reliability. Magnetic modulators of fundamental output are necessarily of rather low impedance and have null stabilities not usually as good as the best transistor chopping circuits. Figure 60, shows the circuit of a "fundamental" type of magnetic modulator. Another limitation is slow response time, which gets slower as sensitivity improves.

Second harmonic magnetic modulators exhibit better performance, probably better than most transistor choppers. The limitations of low impedances and limited frequency response are still present. A typical circuit is shown in Figure 61. Now, we can beat the response time rap by the stabilized amplifier techniques described by Goldberg[8,9] and others,

[6]D-C to A-C Modulators. George Sideris, Electronics, p. 47, January 23, 1959.

[7]Magnetic Amplifier Circuit Applications, describing FERRAC Magnetic Computing Amplifiers, available from AIRPAX ELECTRONICS INCORPORATED, Seminole Division, Fort Lauderdale, Florida.

[8]Stabilization of Wide-Band Direct-Current Amplifiers for Zero and Gain. Goldberg, RCA Review, p. 296, June 1950.

[9]Transistor Operational Amplifiers. Beneteau, Blaser and Lane, IRE Conference, March 1962.

as discussed in later sections. However, it is somewhat illogical to refuse a transistor chopper in a transistor amplifier, and as pretty often repeated in this text, the mechanical chopper may not be too reliable, but 5,000 hours of life is considered normal and much longer periods have been measured.

FIGURE 62.

Magnetic Data Logging Amplifier. Input impedance 50,000 ohms at a gain of 100; variable gain from 100 to 10,000; complete input isolation.

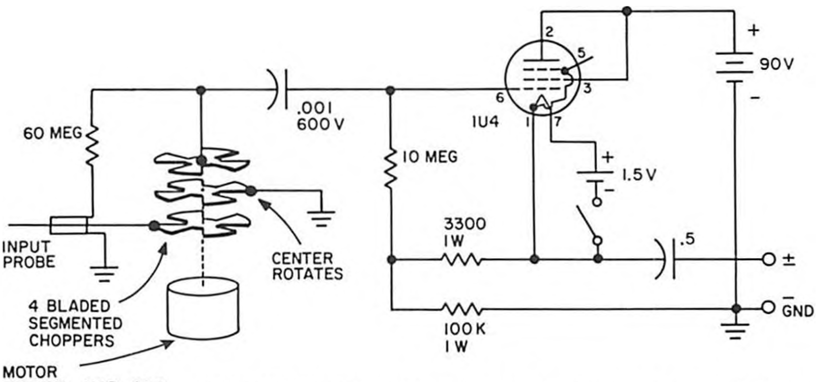


FIGURE 63. *Circuit of electrometer modulator.*

The null stability of magnetic amplifiers can be pretty good, even at moderately high impedances.[10] Figure 62, shows an AIRPAX data amplifier. Measurements of a typical amplifier show nulls within +10 and −20 microvolts over 100 hours and −55°C to +85°C. The specification limit is ±100 microvolts. The input impedance is 50,000 ohms at a gain of 100, reducing to 10,000 at a gain of 500.

CAPACITIVE CHOPPER. A useful device for the investigation of small electrostatic charges may be built as shown in Figure 63. This device is essentially a capacitive "chopper" input circuit followed by a "floating" cathode follower.

The output, which consists of essentially rectangular pulses at 80 cycles/second, gives an indication of input polarity by changing phase. These pulses may be viewed on an oscilloscope. The sensitivity of this device is such that the determination as to the anionic or cathionic nature of a plastic may be determined by rubbing the material against the probe.

[10]A Low-Level, High-Accuracy, D-C Magnetic Amplifier. Blas A. Mazzeo, Electrical Manufacturing, November 1958.

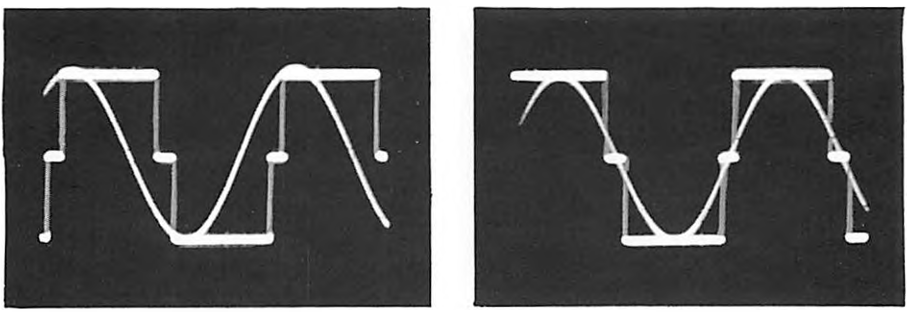
SECTION VII

PHASE RELATIONSHIP

A modulator can usually be a demodulator, a mirror image relationship. If one chopper is the modulator, and another similar but physically separate chopper is the demodulator, it is usually only necessary to parallel the drive coils to obtain close tracking in phase, even over time, temperature and frequency excursions. This fact is not readily understood. In practice, however, one finds that the mechanical linkage of a double pole chopper cannot be perfectly rigid, and that the electrical link is really not much different.

In the frequent application of stabilizing, the Goldberg circuit, a single SPDT chopper both modulates and demodulates. Obviously, in this circuit a knowledge of phase angle is useless. (In fact, this circuit, as later discussed in detail, is so tolerant of poor chopper performance that almost all parameters can change 50% before becoming observable.)

Phase relationship of drive voltage to contacts becomes important when the demodulator, or load, is another phase sensitive device, such as a servo motor or a phase controlled demodulator. Unless this other component bears some accidentally happy phase relationship to the chopper, it will be necessary to introduce phase compensation in the circuit. Obviously, the phase stability with frequency, time, voltage and temperature becomes important.



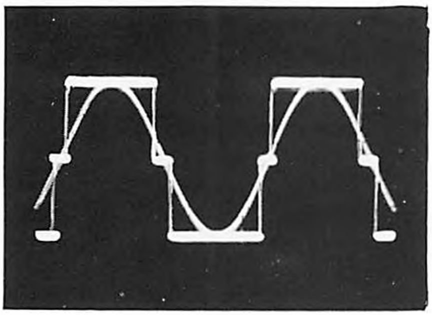
65° LAG 0°

FIGURE 64. *Phase angle relationship to drive can be adjusted readily.*

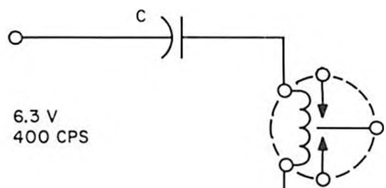
In any mechanical chopper there exists two components of phase lag, the L/R relationship of the drive coil, and an electromechanical lag of the armature. Where chopper phase angles are of interest, 0° and 90° are most popular for obvious reasons, and 0° is the most easily obtained by external capacity. Introduction of series capacity introduces a leading current, and it is quite practical to correct most choppers to zero. Figure 64, illustrates a Type 300 chopper photographed in relationship to its

driving signal (a Type 175 chopper operating at 60 cycles provided the time sharing arrangement to permit the superimposed scope pictures).

At 6.3 volts a simple series capacitor is used to adjust phase, as in Figure 65. The capacitor values are given for several 400 cycle choppers (this is not practical at 60 cycles). At higher voltages a series-parallel *RC* combination becomes preferable, as in Figure 66.



0°



CHOPPER TYPE	C (IN MFD)
300	1.33
360	1.33
370	1.6

FIGURE 65. *Introduction of series capacity permits operation at zero phase angle.*

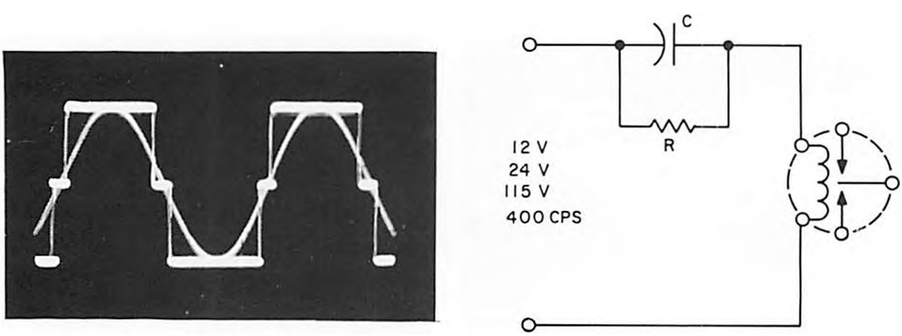


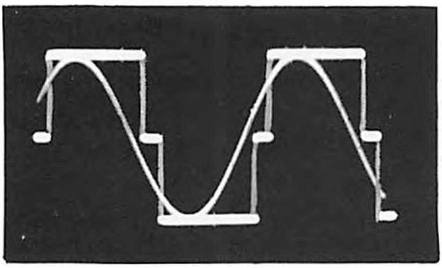
FIGURE 66. *Series-parallel combination provides optimum performance.*

There is an infinity of possible combinations, and optimum values for various drive voltages, such as 115 volts, are given in Figure 67. Now, a zero phase angle is not necessarily the best value, due to other phase shifts which probably exist in amplifiers, etc., but these values will provide a good beginning. They are recommended by the factory as providing best performance.

CHOPPER TYPE	TOTAL PHASE ANGLE (DEGREES)	ELECT. PHASE (DEGREES)	MECH. PHASE (DEGREES)	DRIVE CURRENT MA	COIL Z (OHMS)	115 VOLT NETWORK		24 VOLT NETWORK		12 VOLT NETWORK	
						R OHMS	C mfd	R OHMS	C mfd	R OHMS	C mfd
300	65	30	35	22.5	280	4300	.061	1200	.33	1100	.68
350	65	30	35	24	262	6800	.050	1200	.33	1200	.68
360	55	30	25	22.5	247	4300	.050	1200	.33	1000	.68
370	55	24	31	26	242	5100	.050	1000	.39	840	.94
600	65	32	33	50	126	2600	.10	560	.68	430	1.68

FIGURE 67. *Circuit of Figure 66 permits zero phase angle when RC network uses values listed, at 400 cycles.*

Over wide temperature extremes the copper resistance of a chopper changes considerably. The phase stability is improved by a constant current drive, like that of Figure 68. The phase angle of the chopper will be close to the "mechanical" angle, since compared to the inductive reactance at 400 cycles, the circuit resistance becomes high, the current lag becoming negligibly low.



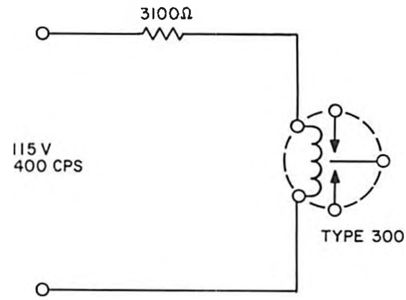
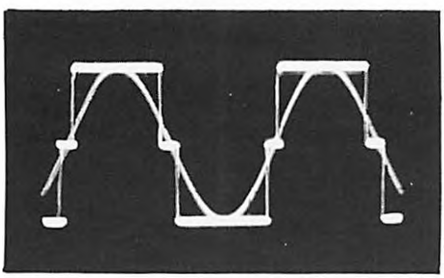
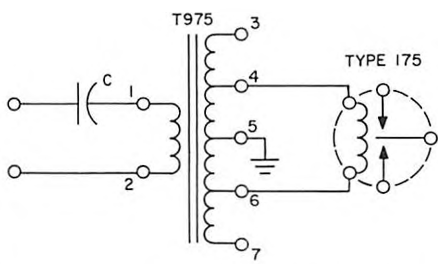


FIGURE 68. *Constant current source produces angle of about 35°.*





	C (IN mfd)	CONNECTIONS
ONE CHOPPER	.250	4-6
TWO CHOPPERS	.540	3-7

FIGURE 69. *Transformer permits zero angle and grounded center tap on drive.*

It may be desirable to shift a phase angle and still permit a balanced drive (to keep noise at a minimum). Figures 69 and 70 describe an AIRPAX Type T975 transformer for use with Type 175 choppers at 60 cycles. The exciting current of the transformer primary affects the angle, as also does the chopper load, and the primary inductance of the transformer becomes important. Type T975 transformers, tested with Type 175 choppers for optimum performance at zero angle, are stocked by AIRPAX at the Cambridge Division.

The phase stability of a Model 30A or 40A series chopper is improved, as compared to larger choppers, since the point of self-resonance is considerably higher. Self-resonance of a Series 300 is roughly 650 cycles, that of a Model 40A may be several thousand cycles. It becomes possible to operate on a flatter portion of the curves.

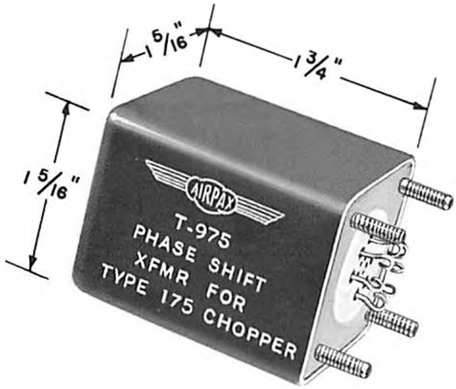


FIGURE 70. *T975 is a small transformer.*

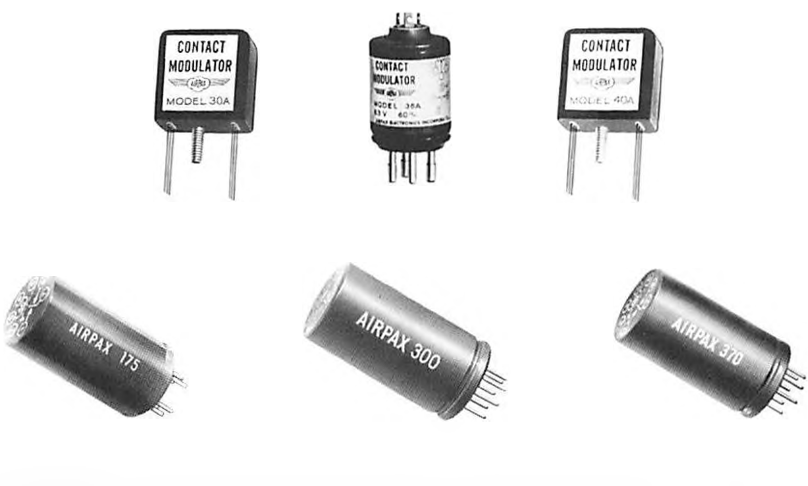


FIGURE 71. *Basic AIRPAX electromechanical chopper types.*

Figure 71 shows the external physical differences of basic AIRPAX chopper types, and Figure 72 illustrates component values for zero correction of Series 30A and 40A. Series 30A and 40A have a special advantage. Being essentially free of noise due to internal coupling to the

drive, they can be operated unbalanced, i.e., with one side of the drive at ground. This does complicate the circuitry layout because noise can easily be externally coupled, as later described under the subject of noise.

MODEL 30A			MODEL 40A		
DRIVE VOLTAGE	NETWORK VALUES	RESULTANT PHASE ANGLE	DRIVE VOLTAGE	NETWORK VALUES	RESULTANT PHASE ANGLE
115	R = 5.3K	20	115	R = 2K	30
115	R = 5.5K C = 0.19μF	0	115	R = 2.6K C = 0.11μF	0
24	R = 1K C = 1.4μF	0	24	R = 520Ω C = 0.7μF	0
12	R = 450Ω C = 5.5μF	0	12	R = 400Ω C = 1.8μF	0
6	C = 21μF	0	6	C = 3.3μF	0

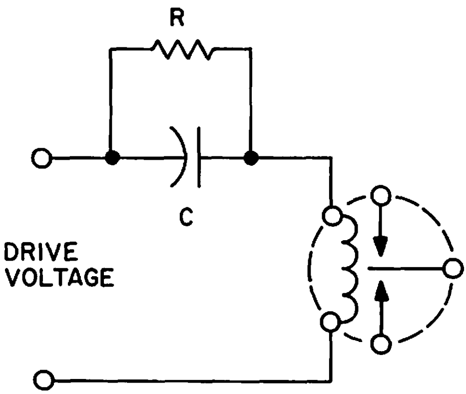


FIGURE 72. *Network values for chopper Models 30A and 40A.*

SECTION VIII

THE AMBIENT CHANGES

Engineers become concerned with component changes under extremes of temperature, altitude, drive frequency, vibration, humidity, shock, dust, fungi growth, and other deleterious assaults of man and nature. Being hermetically sealed in metal and glass, an AIRPAX chopper is resistant to altitude, dust, fungus and humidity.

Humidity could, during prolonged exposure and while carrying substantial DC polarization, cause electrolytic tracking across glass. This is an old story in transformer feed-through leads; fortunately choppers seldom carry sufficient DC potential to reduce the insulation resistance of glass enclosed terminals appreciably.

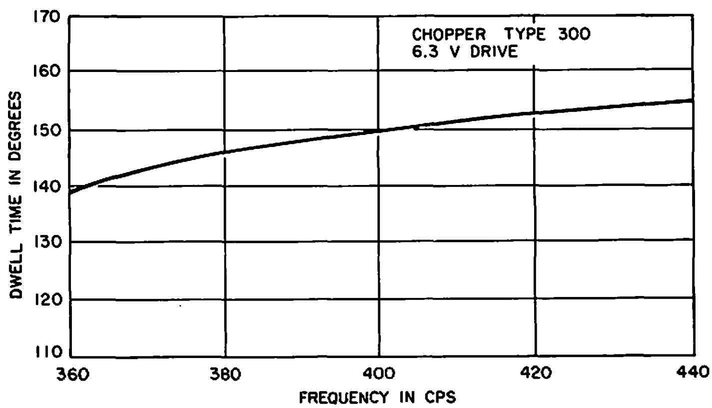


FIGURE 73. *Variation of dwell time with driving frequency.*

The two parameters of a chopper most likely to be affected are dwell time and phase angle. Noise is not usually changed, again due to hermetic sealing. It should be noted that contamination and moisture on the socket or external connections are quite likely to cause noise, substantial values, in fact, from the presence of electro-chemical potentials.

Dwell time is frequently important, as a change in dwell time will be effectively a change in gain in some circuits. Phase angle is not important in many DC amplifiers, except for the relation between modulator and demodulator. (Two AIRPAX choppers of the same part number will remain almost identical with each other, even while they change from a change of environment. Several users have taken advantage of this fact to provide as high as 4-pole double throw operation, by simply operating units in parallel.)

ENVIRONMENT AND THE DWELL TIME. Figure 73, illustrates the change in dwell time with driving frequency of a production Type

300 chopper. Specification limits are usually 380 and 420 cycles. The dwell time increases with frequency, but the chopper remains balanced, which is to be expected if the dwell times are initially symmetrical. Figure 74, is another typical curve, the effect of varying the coil voltage from 3 to 9 volts (5.9 to 6.7 is a more usual range to be expected). Figure 75, illustrates the relative indifference of the chopper to tempera-

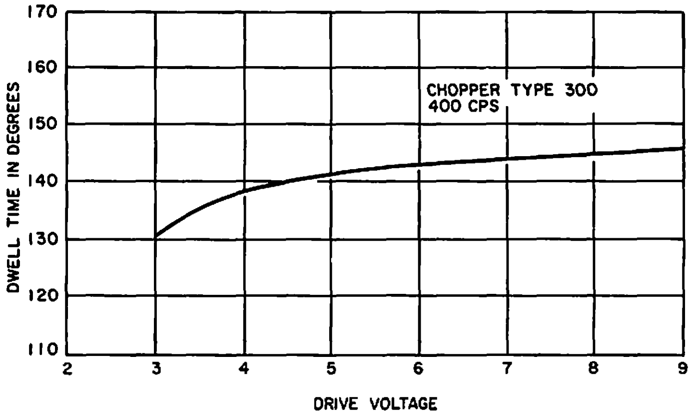


FIGURE 74. *Variation of dwell time with driving voltage.*

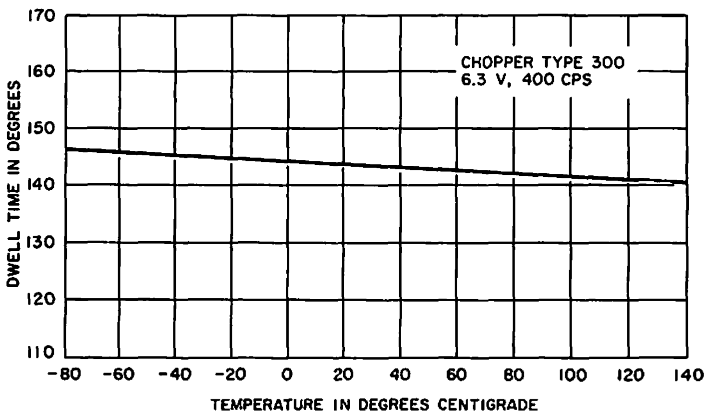


FIGURE 75. *Temperature effect on Type 300 chopper.*

ture, over the extreme of $-65°C$ to $+125°C$. Again specification limits are less than the capability of the chopper, which normally operates from $-50°C$ to $+85°C$.

Figures 73, 74, and 75, are values averaged from large groups, accordingly are illustrative of what can be expected at 400 cycles, at least from an AIRPAX Type 300 chopper. Figure 76, is the measured dwell time during 25,000 hours of life testing (2.86 years of operation). The

probable spread of dwell time figures is well illustrated in Figure 77 showing the extreme range of dwell time as measured on 20 Type 300 choppers during 1,500 hours of life testing. Complete measurements were made every 250 hours and the dwell time was recorded automatically every few hours.

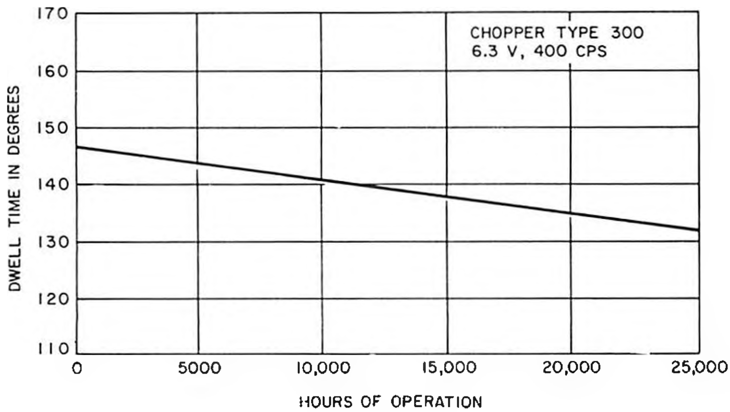


FIGURE 76. *LIFE TEST DATA. Chopper operating continuously in "dry" circuit. Dwell time recorded every 6 hours; a complete test performed every 100 hours.*

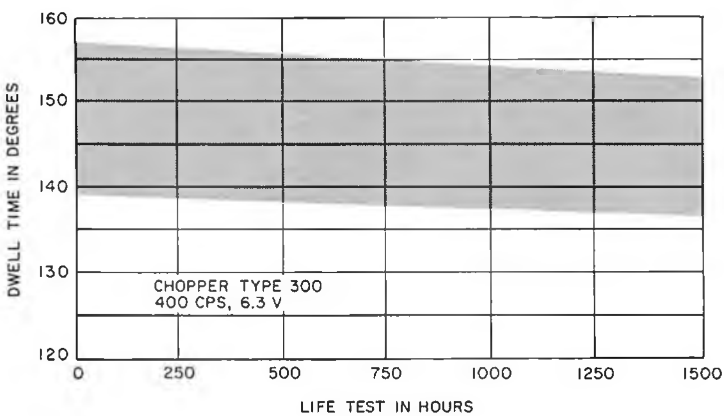


FIGURE 77. *Maximum and minimum dwell times recorded on 20 Type 300 choppers selected at random from production.*

THE PHASE ANGLE VARIATION WITH AMBIENT. The phase lag of a signal from chopper contacts is composed of two elements, the *L/R* relationship of the coil plus a "mechanical" lag of the armature. As the operating frequency increases the phase angle also increases, and

the relationship is plotted in Figure 78. An increasing voltage reduces the angle, as illustrated in Figure 79. When operated from a ship-borne alternator, the increase in voltage and frequency may occur together; if so there will be a tendency to neutralize each other. While the plot is given over 360 to 440 cycles, and 3 to 9 volts, a chopper is not often required to operate over such extremes, as discussed earlier. When operated over a wide range of ambient temperature a phase angle change,

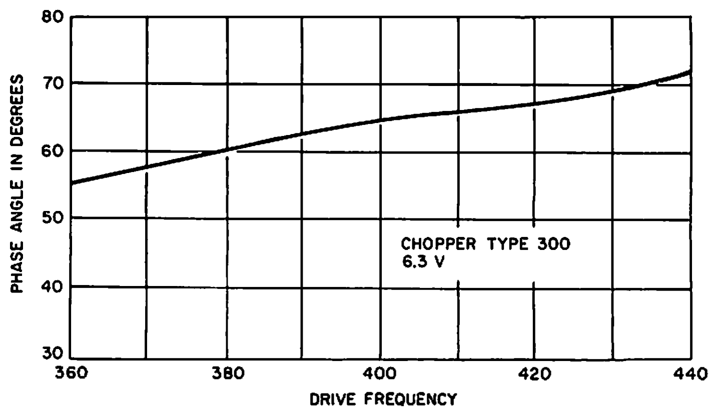


FIGURE 78. *Phase angle variation with driving frequency.*

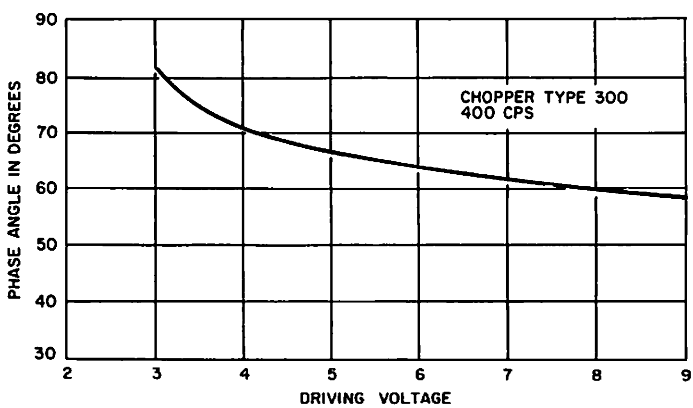


FIGURE 79. *Effect of driving voltage change.*

Figure 80, occurs which is, to a large extent, due to the change in coil resistance. Between −65°C and +125°C the coefficient of copper will vary the resistance of a Type 300 chopper from 110 ohms to 236 ohms. Since the reactance of a Type 300 is about 150 ohms at 400 cycles, the effect of temperature on the coil time constant will be appreciable.

Uniformity of production choppers is illustrated by Figure 81, a plot of the distribution observed on 100 production choppers.

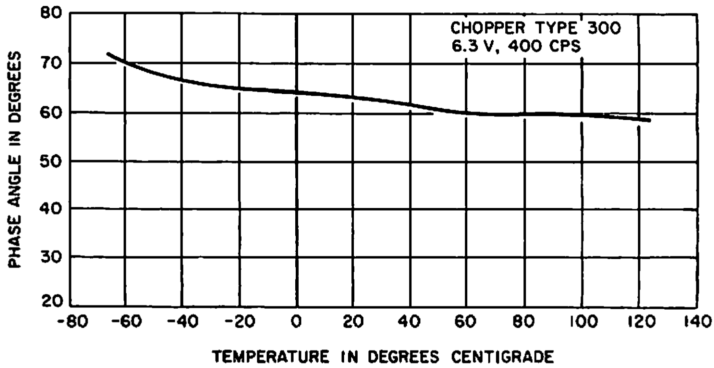


FIGURE 80. *Phase angle change with temperature is largely due to change in coil resistance.*

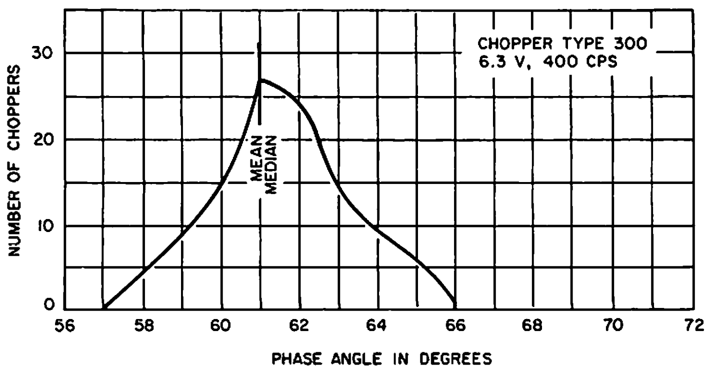


FIGURE 81. *Distribution of phase angle, 100 production choppers.*

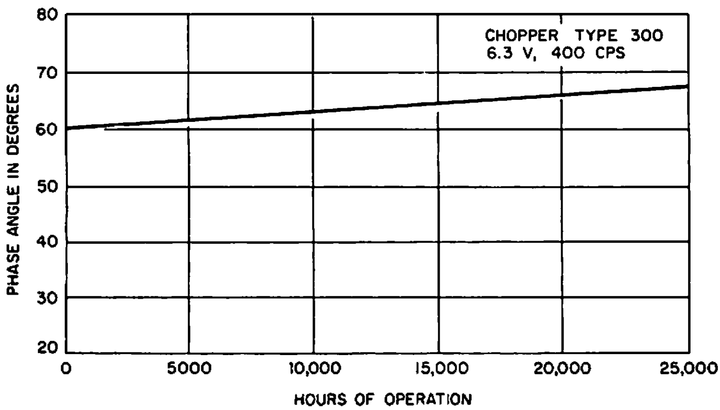


FIGURE 82. *When operated in a dry circuit a Type 300 chopper maintains a nearly constant phase angle.*

When the phase angle becomes important, the effect of wear and changes during life become important. Figure 82, shows the almost negligible changes during practically three years of continuous "dry circuit" operation of a Type 300 chopper at 400 cycles.

Some of the phase adjusting networks will provide phase stability over a very wide range of operating frequency, as plotted in Figure 83.

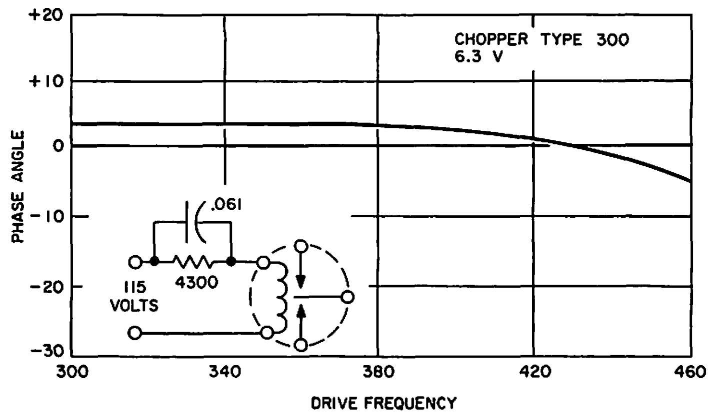


FIGURE 83. *The use of networks improves the phase stability.*

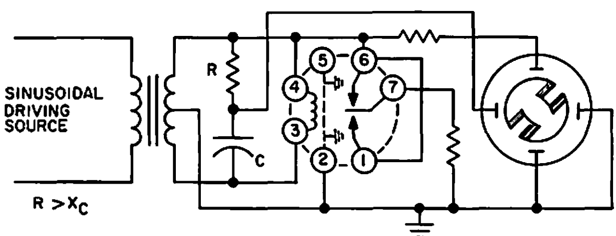


FIGURE 84. *Contact derangement as it appears on a polar scope presentation.*

VIBRATION PROBLEMS. Chopper Type 300 and many others, employ an armature hinged at one end. If the acceleration forces applied to a chopper are sufficiently high, there will result what is termed contact "derangement," essentially a modulation of the phase angle.

The effect of vibration on chopper action is measured in electrical degrees of contact derangement. Contact derangement is the aggregate fluctuation in make or break due to bounce, chatter, phase modulation, and dissymmetry due to vibration. Figure 84, shows these characteristics on polar coordinates as they appear in a simple test circuit.

Type 300 will survive vibration testing without damage, but as noted in Figure 85, at frequencies approaching 400 cycles, the acceleration forces exceed the armature drive forces. Mechanical self-resonance occurs at about 550 cycles, at vibration frequencies in this vicinity the acceleration forces take over and the armature ceases to follow the drive signal.

In Series 370 choppers, the armature is supported on a pivot. As a result, external mechanical vibration imparts only translation to the armature — the same as it does to the fixed contacts. The balanced structure effectively isolates the armature from extraneous torques that could interfere with its desired motion or displace it relative to the fixed contacts.

The drive coil and polarized magnetic circuit couple to the balanced armature at one end. The moving contact is mounted at the opposite end, well removed from the magnetic field to minimize stray pickup.

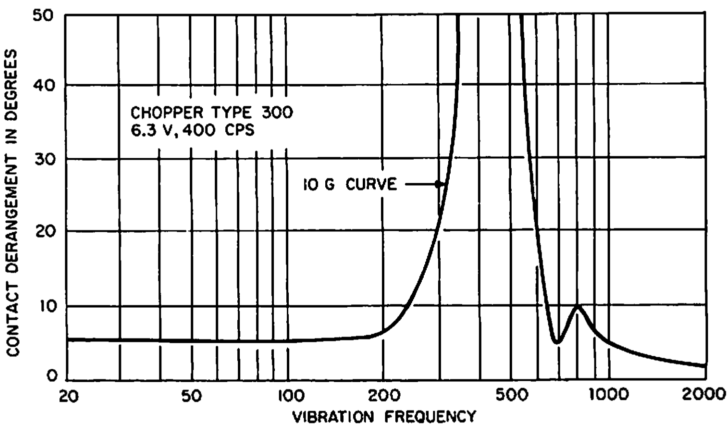


FIGURE 85. *Type 300 choppers are not damaged by high frequency vibration but do not operate correctly.*

Because unwanted vibrations act on the armature at its center of gyration, they produce no rotation, while the drive field acts at the end of the armature to develop maximum torque — leading to an efficient drive mechanism that requires low drive power.

Figure 86, shows that the effect of external vibration is isolated by this balanced structure. The curve, from a typical Type 370 chopper, shows contact derangement as a function of frequency for vibrations of constant 15 G in the plane that is most sensitive to vibration. The natural resonance of the armature is seen to be about 500 cps. Even here, however, external vibration has little effect on chopper action. The curve was obtained with the 400 cps voltage to the drive coil at 5.7 rms volts; that is, with the driving force at the lower limit of its specified tolerance. This curve shows what can be expected under the most adverse conditions.

Figure 86, examined the worst possible conditions for a Type 370 chopper, using the plane of maximum contact derangement. Figure 87, is similar data, but taken at right angles to the direction of Figure 86, and its negligible derangement will be obvious. Most AIRPAX miniature choppers having a 7-pin base, whether with solder terminals or plug in, have their internal works aligned with the illustration of Figure 90. Plane 1 is the direction of contact motion, plane 2 at right angles, plane 3 is of course parallel to the frame, and the chopper can.

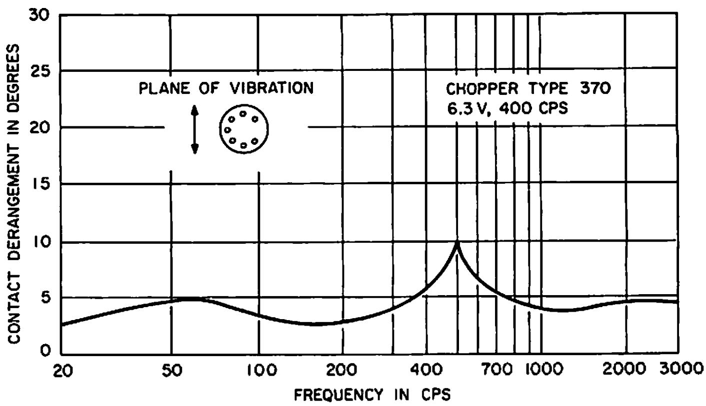


FIGURE 86. *Effect of 15 G vibration on a Type 370 chopper.*

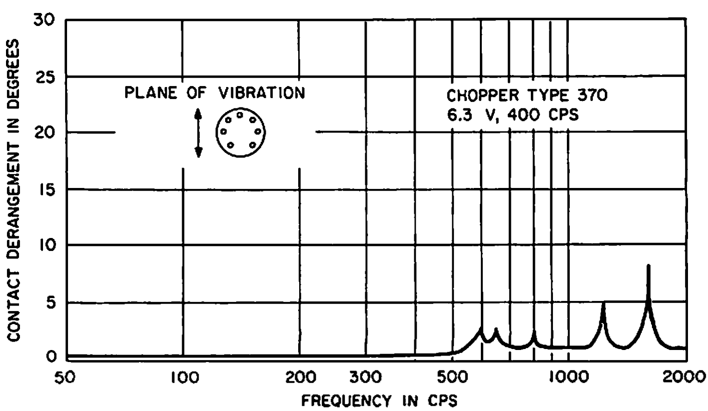


FIGURE 87. *Type 370 chopper vibrated in lesser plane.*

LOW FREQUENCY CHOPPERS. The attack of a changing environment has been confined up to this point to 400 cycle choppers. At power frequencies of 50 or 60 cycles, components are often not asked to withstand the degree of change found in aircraft or missile equipment. However, the 60 cycle chopper can be built to withstand extremes as readily as at aircraft supply frequencies.

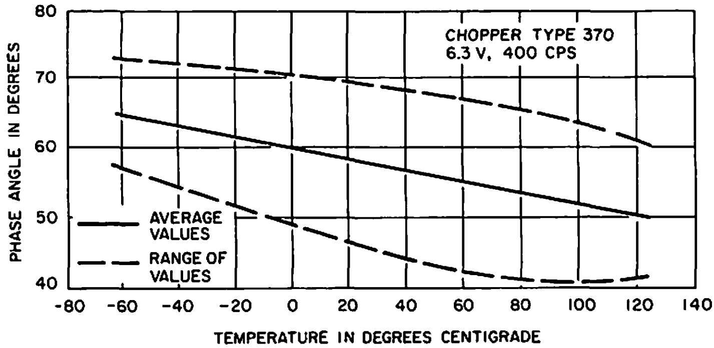


FIGURE 88. *Phase versus temperature of Type 370.*

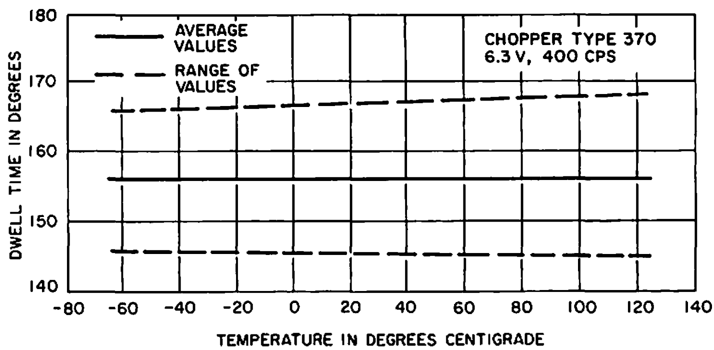


FIGURE 89. *Dwell versus temperature of Type 370.*

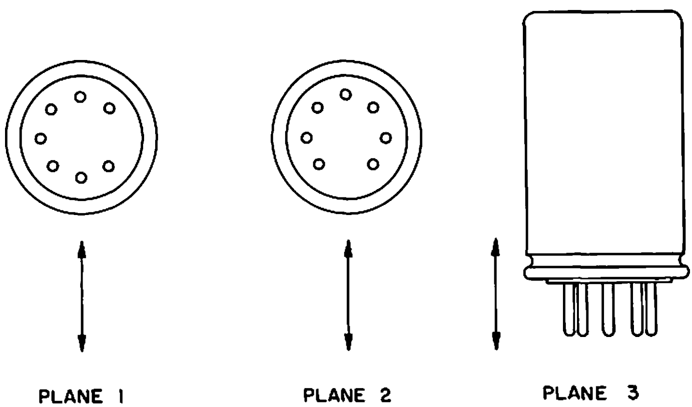


FIGURE 90. *Plane 1 delivers vibration parallel to the direction of contact motion.*

Figures 91 and 92 describe the limits of a family of curves taken on choppers Type 175, when operated at 6.3 volts, 60 cycles, and subjected to an ambient varying from $-40°C$ to $+100°C$. It should perhaps be repeated that individual units do not fall in a random fashion within these boundaries — the upper and lower limits can be considered as describing two individual choppers out of a large group from which this data was derived.

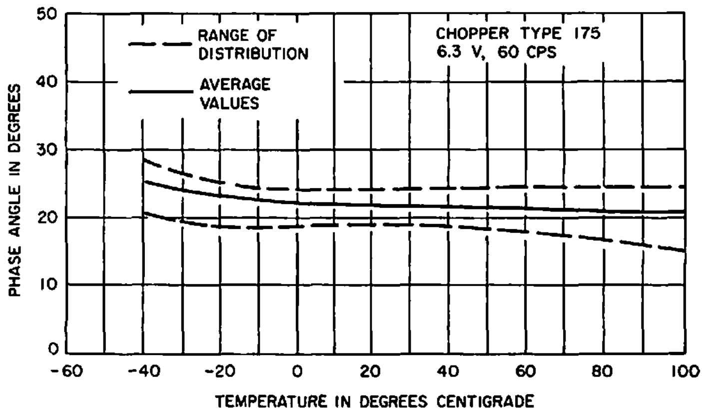


FIGURE 91. *Phase angle spread of a large group of Type 175 choppers.*

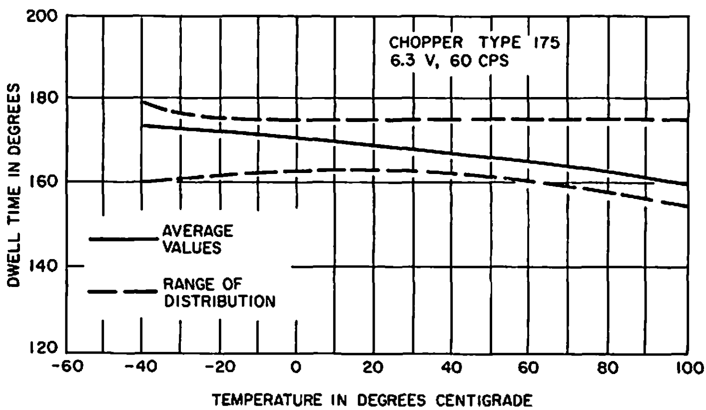


FIGURE 92. *Variation of dwell time with temperature and with production, Type 175 choppers.*

The effect of life up to 10,000 hours is shown in Figures 93 and 94. The units operated in "dry" circuit conditions, except for periodic recording of dwell time (every six hours with automatic equipment). During the several minutes on the recorder, 50 millivolts at 50 microamperes appear across the contacts, not enough to "clean up" the contact surface.

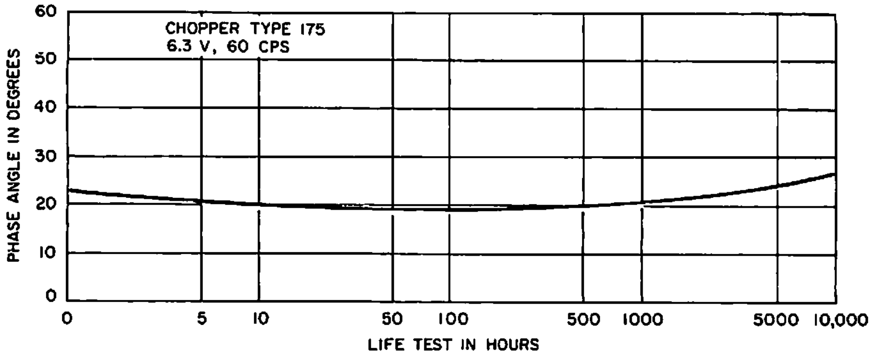


FIGURE 93. *Change of phase angle during 10,000 hours of operation, Type 175 chopper.*

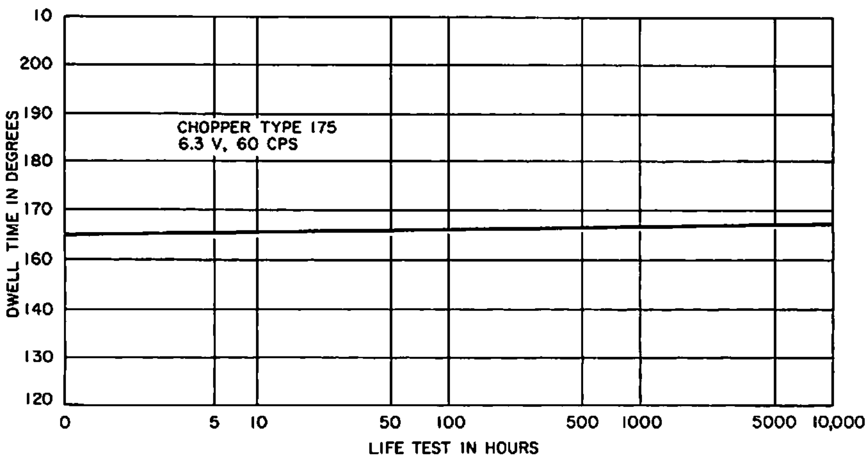


FIGURE 94. *Dwell time of a Type 175 chopper operated for 10,000 hours.*

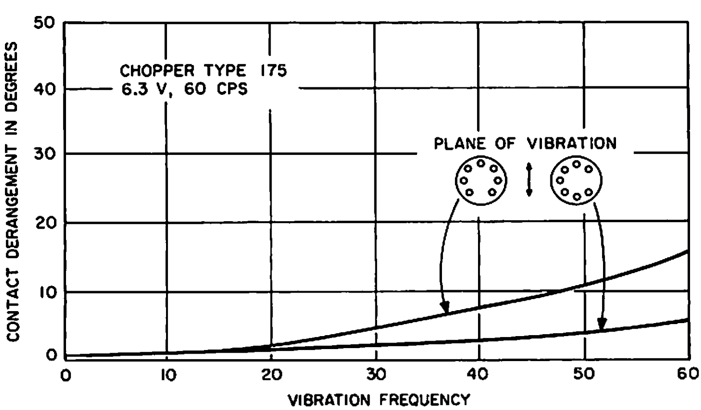


FIGURE 95. *Effect of vibration on a Type 175 chopper.*

As the choppers were operating satisfactorily at the conclusion of this particular test, it is difficult to predict how long they might last — apparently they will continue indefinitely.

Figure 95, shows the amount of contact derangement from a vibra-

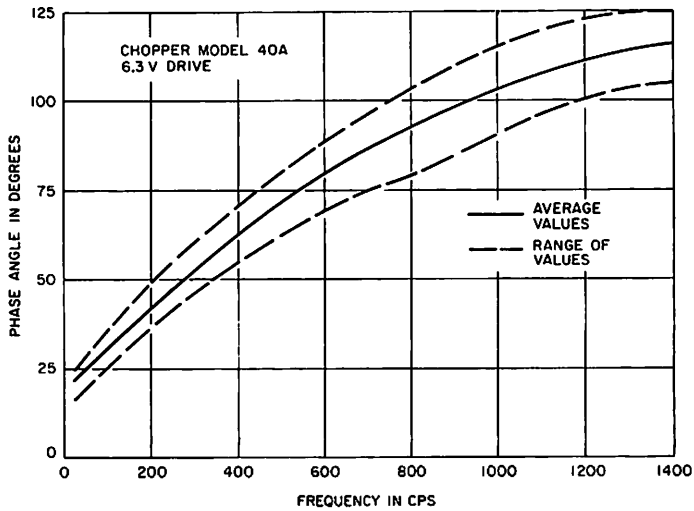


FIGURE 96. *Phase change of Model 40A with frequency.*

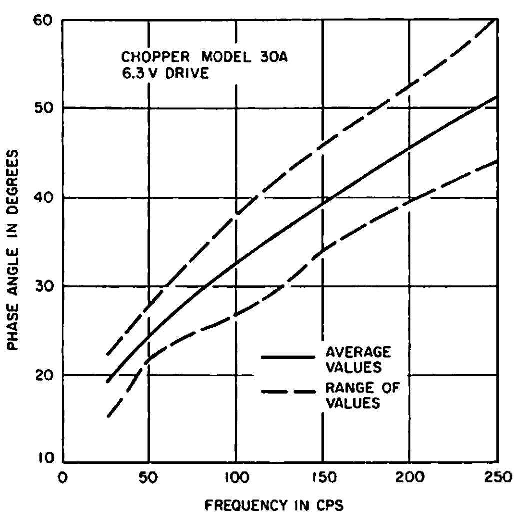


FIGURE 97. *Phase change of Model 30A with frequency.*

tion table excursion of 0.05 inches. Planes 1 and 2 of Figure 90 are plotted, being the worst and next worst planes. The contact derangement is usually symmetrical, appearing as a symmetrical modulation of the phase angle. Being symmetrical, it often does not cause offset in a DC

amplifier, and as a 60 cycle chopper amplifier pass-band will be about 15 cycles, the vibration component may not appear in the output. In any case the chopper is undamaged. Just as shown earlier for the Type 300, Figure 85, at high vibration frequencies the phase modulation becomes total, i.e., the chopper will follow the vibration.

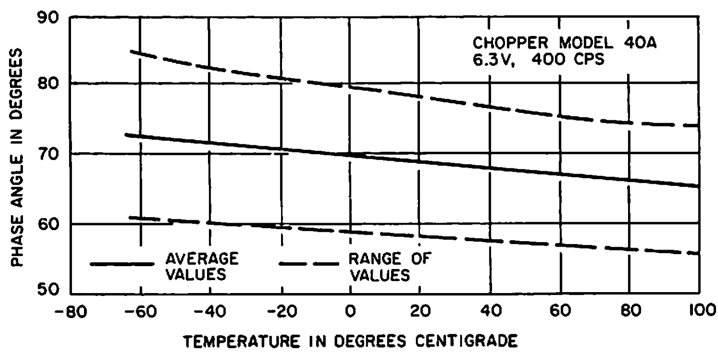


FIGURE 98. *Phase change of Model 40A with temperature.*

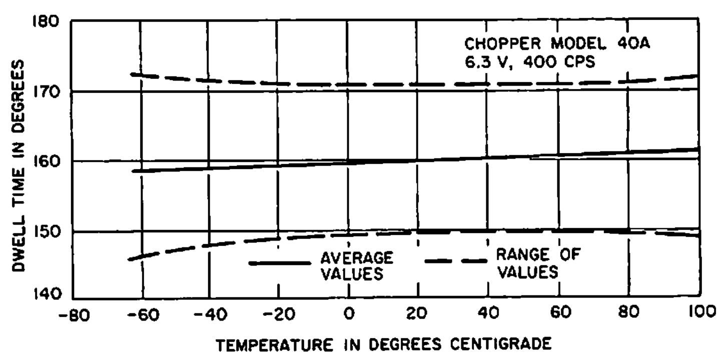


FIGURE 99. *Dwell time of Model 40A versus temperature.*

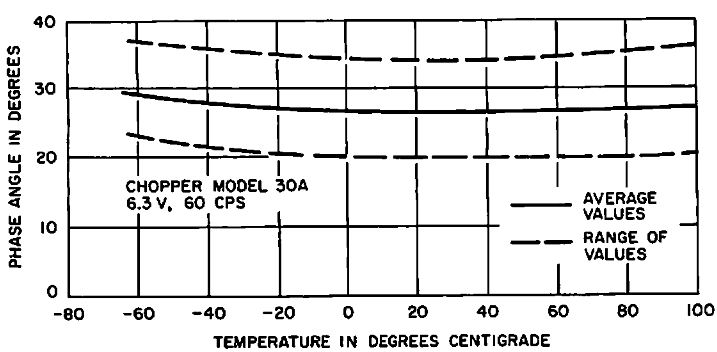


FIGURE 100. *Phase change of Model 30A versus temperature.*

Types 300 and 370, at 400 cycles, and Types 175, at 60 cycles, can only be good examples of characteristics observed on a few popular chopper models. Obviously other designs will have different response to ambient variations, and the user should ask for data on the model he finds most suitable for his application.

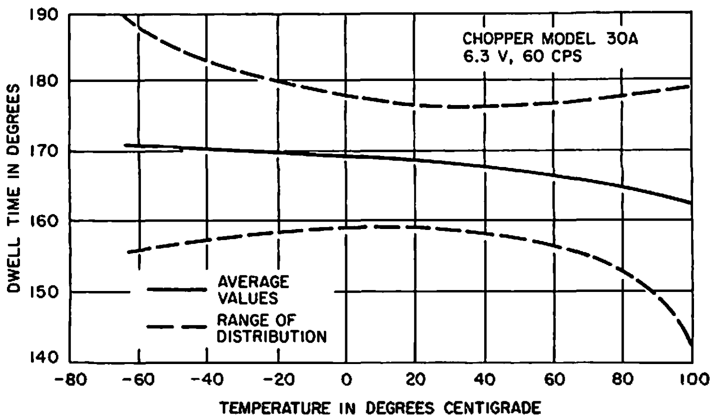


FIGURE 101. *Dwell time of Model 30A versus temperature.*

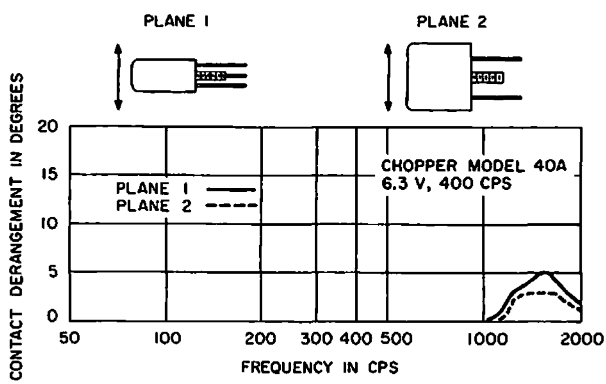
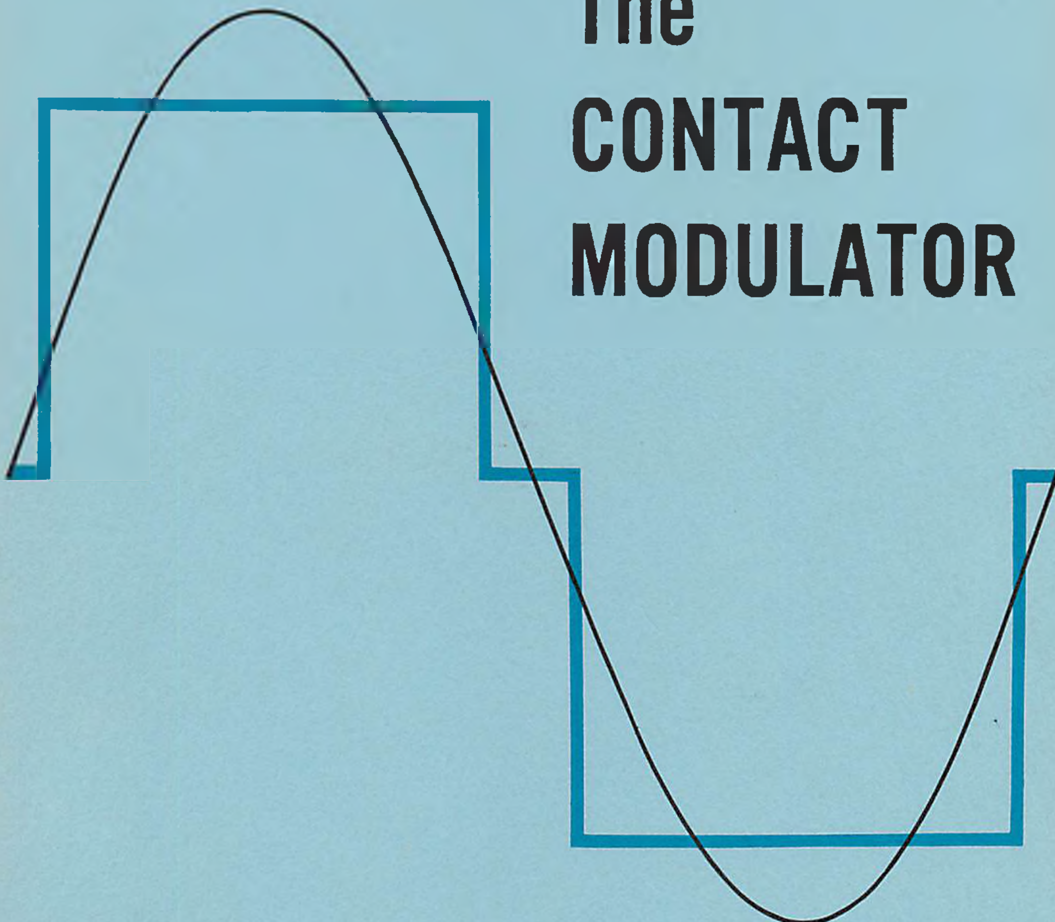


FIGURE 102. *Contact derangement of Model 40A under vibration.*

The "micro" series of choppers, of the postage stamp variety, has considerably different performance under environmental changes. As a general rule they are somewhat more stable. The performance to be expected can be observed in the data presented in Figures 96 to 102 inclusive.

AIRPAX

The CONTACT MODULATOR



PART 3: Modulation Methods

THE THIRD OF A SERIES

AIRPAX

THE CONTACT MODULATOR

PART III

MODULATION METHODS

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TABLE OF CONTENTS

SECTION I

SOME BASIC PRINCIPLES

Mechanical choppers were originally developed for modulating very low level DC signals which could not be amplified accurately in any other way, and this is still their prime function. Many other applications have since been found such as demodulation, time-sharing and DC isolation, to quote three examples, but modulation is still the most important.

Any device, which in the broadest sense will multiply, can be used as a modulator, and several have been outlined in Part 1 of this series. The mechanical switch, however, is especially efficient since it can multiply a signal by very nearly zero or very nearly unity — as wide a range as anyone could wish for — and this is because its open resistance is very high, its closed resistance very low, and its noise (in a well designed unit) extremely small. This section deals with modulators based on the use of switches. Transistors also make quite good switches, and providing their limitations are borne in mind, many of the following remarks apply to circuits using them as well.

Switch-type modulators convert a direct input signal into a nominally rectangular AC carrier wave at the chopping frequency. This "square wave" can then be amplified by a relatively simple AC-coupled amplifier whose DC drift does not matter, and reconverted to DC at a higher level by a demodulator and filter. If the input changes polarity, the carrier reverses its phase and the demodulated output changes polarity as well. The response of such a system to alternating or step inputs depends on the design of the modulator and the demodulator. If both are full-wave and accurately synchronized, for example, and the off time is short, the input signal frequency may approach that of the carrier. Signals above carrier frequency are also transmitted faithfully during the on times. In many practical circuits, however, the use of half wave circuits and the possibility of some drift in contact adjustment make it desirable to smooth the output, and this will limit the response to some fraction of the carrier frequency. Although some sort of demodulator is invariably necessary to extract the information from the carrier, it need not have an electrical output. For example, an AC servo motor is a demodulator whose output is angular velocity. One of its inputs is the AC carrier, and the other is an AC reference of the same frequency.

When the output of a chopper amplifier is zero, the input should also be zero, but in practice it invariably has a small value termed offset, zero offset, or drift. The magnitude of this offset is perhaps the most important property of the system, since it determines the overall accuracy. Strictly speaking, both stable and unstable components are present, the former responding to compensation, but for present purposes there is no need to

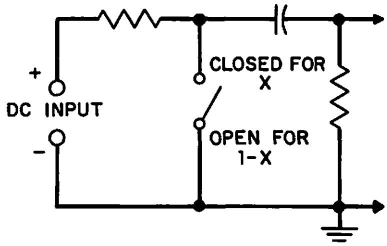


FIGURE 1.
On-off modulator.

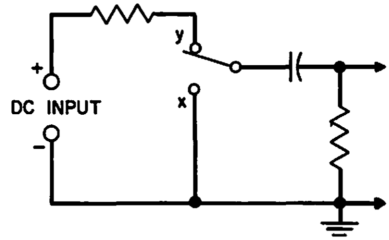


FIGURE 3. *SPDT modulator (BBM).*

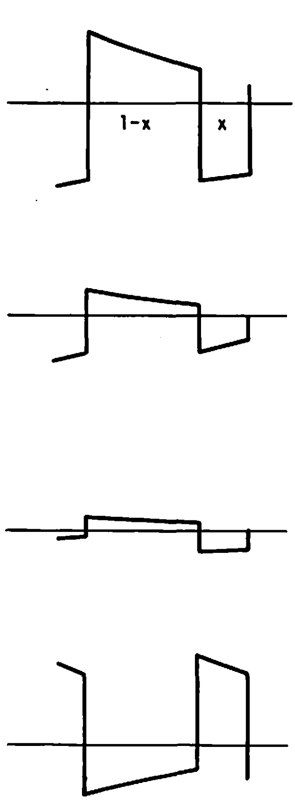


FIGURE 2. *Output waveforms of on-off modulator.*

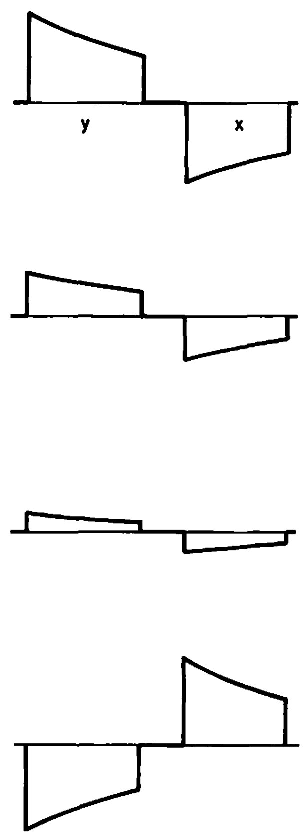


FIGURE 4. *Output waveforms of SPDT modulator.*

distinguish between them. Offset depends on many factors such as pickup in the signal, thermal e.m.f.'s in the wiring, amplifier hum, etc., and one test of a good design is the extent to which these effects have been reduced. In the last resort, the offset usually depends on the noise in the chopper itself, values in the order of one microvolt and 10^{-11} amp. being achievable with commercial designs. This is very much better than any direct coupled amplifier can do.

Although the drift can be made very low, the gain stability of the modulator — AC amplifier — demodulator system is not particularly good (except in a few special circuits), being dependent on waveform mark-space ratio. For this reason, negative DC feedback is almost invariably applied around the whole system to stabilize the gain. The modulator then handles a difference or error signal which is kept very small by the action of the feedback. The forward gain is made so large that full output is obtained for an error signal which is no more than e.g. 1% or 0.1% of the input signal. If the gain is increased, the accuracy improves in proportion until the error signal has been reduced to the same level as the modulator offset. Any further increase in gain gives no advantage and renders the system less stable.

The error signal may be derived first and then applied to the modulator as a single input, or alternately the modulator may itself respond to the difference of two large signals applied independently. Very often one of the input signals is fed through a low-pass filter to remove unwanted pickup or noise. The feedback from a demodulator, however, is usually smoothed by a single RC stage to ensure stability of the feedback loop, and it is therefore applied without additional filtering which would provide a second lag.

A simple modulator circuit is shown in Figure 1, and provides a nominally rectangular waveform which may be symmetrical or not as desired. Typical waveforms are illustrated in Figure 2, which also shows how the output changes phase as the input changes polarity, and is zero when the input is zero. Figures 3 and 4, show a rather more efficient circuit with a somewhat different waveform.

Practical waveforms differ somewhat from the ideal, and there is usually some droop, as shown, due to the finite time constants involved. But it can be made small, and we shall ignore it. The rise of leading edges, and the fall of trailing edges can also be delayed if the load contains significant shunt capacity, or the amplifier response is poor, but again this is a secondary and usually avoidable effect and we shall ignore it too.

SECTION II

PRACTICAL MODULATOR CIRCUITS

So much for the basic principles. What is required from a practical modulator, and how is it best obtained? Here are some of the more important factors that may need to be taken into account.

DESIGN CONSIDERATIONS.

1. Need for isolation between DC input and AC output.
2. Need to compare two large inputs.
3. Need to compare two inputs at different potential levels.
4. Degree of immunity required to pickup, both series and common mode.
5. Effects of amplifier noise, hum and low frequency drift.
6. Stability considerations when there is overall DC feedback.
7. Source resistance and amplifier input impedance.
8. Output waveform desired.
9. Type of chopper available.
10. Conversion efficiency.

It will be interesting to look at some practical types of circuits, and see which of these possible requirements they meet, and where they are inadequate.

SIMPLE GROUNDED CIRCUITS. The circuit of Figure 1, is a useful one when the input is initially floating, can be grounded on one side to the amplifier, and does not contain too much pickup. It is economical in contacts, and allows the other half of an SPDT chopper to be used to demodulate the amplified signal, or to act as modulator in a second amplifier. One or other of these arrangements is common in the subsidiary chopper amplifiers used to stabilize direct-coupled amplifiers for analog computer use. The circuit is not particularly efficient, since input current is drawn to no useful purpose every time the switch closes, and for this reason an on-to-off ratio less than unity is sometimes used.

More efficient is the SPDT circuit of Figure 3, which eliminates the wasteful drain by using another contact. Efficiency is not usually a vital consideration however, unless the limits of noise are being approached. Of greater importance is the fact that both these circuits are very susceptible to pickup at the chopping frequency, this being at least partially passed on to the amplifier and so causing a DC offset. Thus both are likely to be unsuitable where there are long input lines, such as occur in thermocouple temperature measurement, unless the chopping frequency differs from that of the supply mains.

COMPARING TWO INPUTS. There are in principle two ways of comparing two large signals so that the modulator output depends on their difference.

The simplest and most accurate is to connect the two sources in series and apply the difference as a single input to circuits such as those in Figures

1 and 3. In this case one of the inputs is operated at a varying potential to ground, assuming the load to be grounded. However, a slight modification (Figures 5a and 5b), in the connection of the load resistor to the DC carrying loop of the circuit permits both inputs to operate with one side at ground potential. When the two inputs are equal, the coupling capacitor becomes charged to the common potential and there is no AC output. Either input can be filtered.

In the other method, shown in Figure 6, the two inputs must be of opposite polarity to ground, in contrast to the above, and are summed as

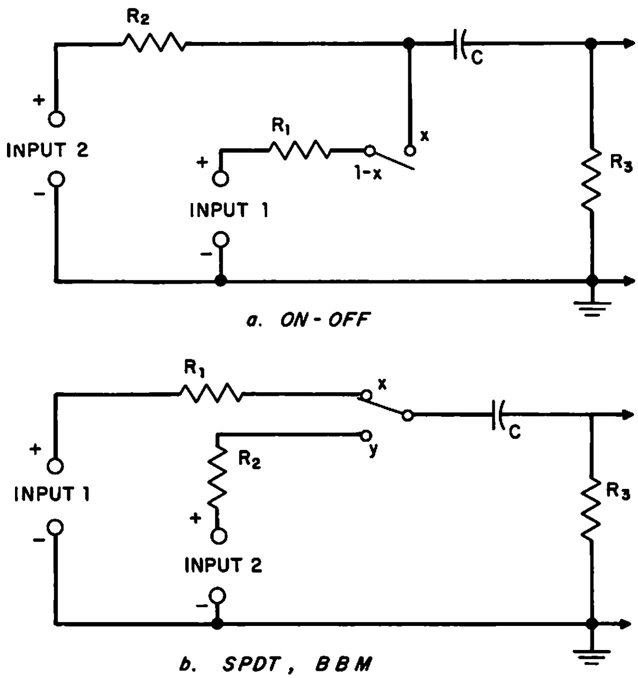


FIGURE 5. *Two grounded inputs can be compared in series.*

currents through two accurate resistors onto one point. Sources of low or known resistance must be used, and the accuracy of comparison depends on that of the resistors.

A CAUSE OF LOOP INSTABILITY. The circuits shown in Figure 5, may be a subtle cause of instability in cases where one of the inputs is in fact a feedback signal derived from the amplified and demodulated output.

Consider either circuit and assume that both inputs are suddenly raised by the same amount. Since the circuit time constants are several cycles long, the coupling capacitor takes a few cycles to charge up, and the output signal meanwhile takes one of the forms shown in Figure 7.

Both signals contain an AC component and a change in DC level, and their effect depends on the design of the demodulator. A balanced demodulator will respond mainly to the AC component, but a one-sided type may respond to the DC shift as well.

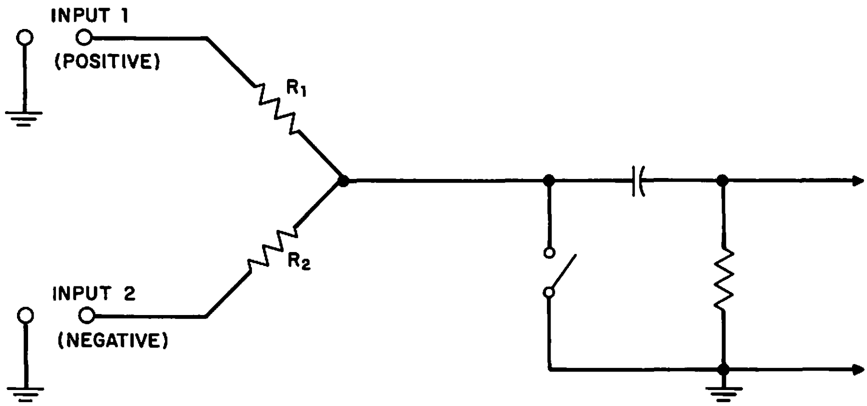


FIGURE 6. *Two inputs can be summed through accurate resistors.*

The essential requirement, however, is that for stability the total effect should be to reduce the feedback signal. Any tendency to increase it is equivalent to a lag in the feedback, and will tend to cause oscillation. Thus in the circuit of Figure 5a, if the demodulator recognizes only the AC

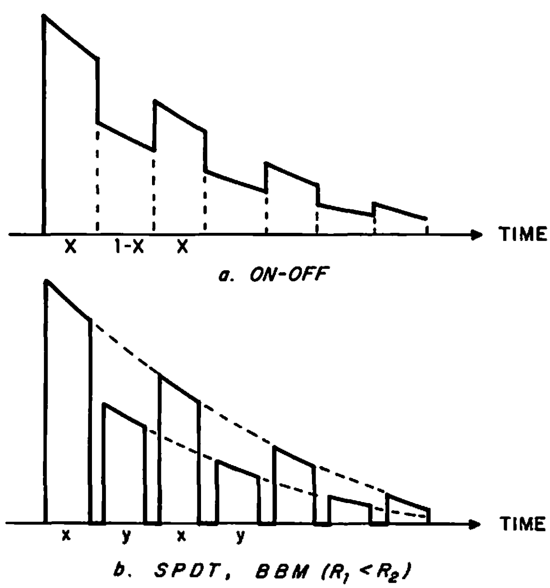


FIGURE 7. *Output due to equal step change in both inputs to the circuits of* FIGURE 5.

component, the feedback signal should be introduced at input No. 1, and R_1 should preferably be small compared with R_2. The AC signal then has the phase associated with too large a feedback signal, and so tends to reduce it. Injecting the feedback as input No. 2 in this circuit is bound to give the effect of a lag. In the SPDT circuit, the resistor associated with the feedback input should be the lower of the two.

The difficulty is avoided if the two inputs are compared first, either in series or by parallel summing through resistors, and then fed in as one input, provided that no filtering is employed.

An ideal arrangement, but a more expensive one, is to use a fully

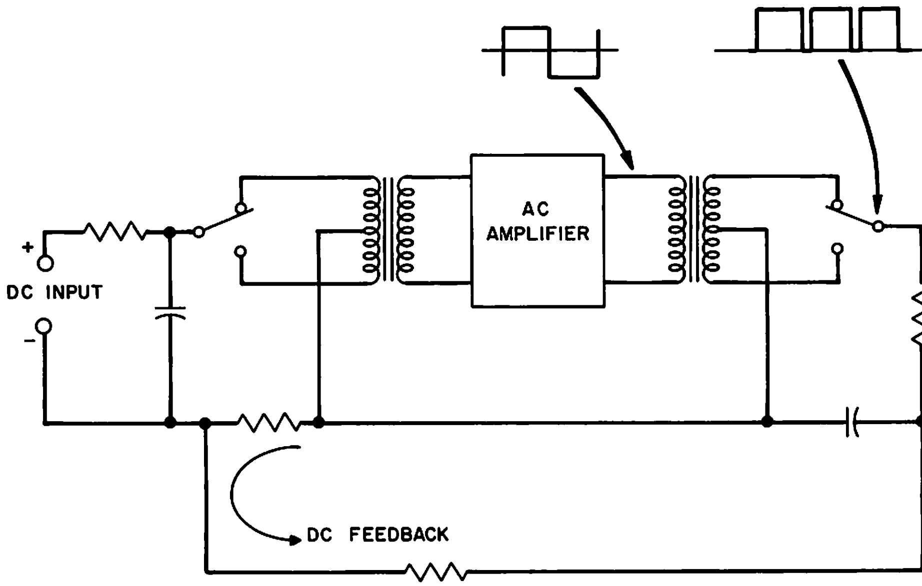


FIGURE 8. *Fully balanced system minimizes instability problems.*

balanced modulator such as the SPDT transformer-coupled circuit shown in Figure 8. The amplifier then handles pure AC, with no changes in DC level even during transient conditions. Use of a balanced demodulator is an added safeguard against unbalance developing in the modulator. It also eases the smoothing problem, and it is worth noting that, in the ideal case, with perfectly synchronized demodulation, zero transit time in the switches, and no droop in the waveform, no smoothing is required at all.

PICKUP. Pickup at the chopping frequency can be a very considerable problem, either by causing a DC offset, or by blocking the subsequent amplifier. There are two distinct forms.

1. A series voltage may be induced by stray alternating magnetic fields if the input circuit is not loop-free. It can be reduced by a single or multi-stage filter on the relevant input, the amount of filtering tolerable depending on the response required from the modulator. If the filter is a balanced one, as in Figures 9 and 10, theoretically none of the fundamental (e.g. 60 CPS) frequency is passed on to the amplifier, and in practice with mechanical choppers an improvement of 10 or more times can be anticipated. However, the 120 CPS and higher harmonics transmitted can still block the amplifier, and some designs of amplifier include a filter to reject all but the carrier frequency.

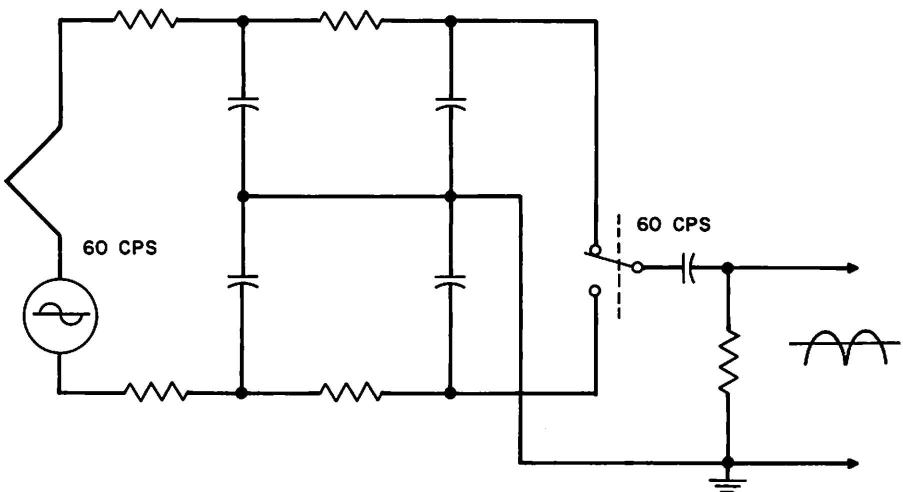


FIGURE 9. *Balanced circuit converts series pickup to double frequency.*

2. A current may be fed to the input circuit as a whole through stray capacities from e.g. supply mains. As it flows to ground through the circuit, voltages are developed which may or may not reach the amplifier according to the type of filter. The circuit of Figure 9, is not balanced against this common mode effect, but does attenuate it. The transformer-coupled circuit of Figure 10, with balanced filter, is entirely proof against it, but in this case another effect comes into play. The primary winding as a whole has an AC voltage to ground and this acts through the inter-winding capacity to transmit a certain amount of the fundamental. This effect can be a serious one, and the inter-winding screen must often be very good. If it is perfect, of course, this effect vanishes.

TRANSFORMER COUPLING. Transformer-coupled circuits are used when it is desired to isolate the DC input from the AC output. In practice this means that the input is either connected to some circuit which is not at ground potential, or is tied to some distant ground point which cannot be assumed to be at the same level as the amplifier ground. They do not isolate one DC input from another, however, and any feedback signal must also be able to take up the input potential. One way of providing this is to use a motor-driven slidewire fed from a dry battery or floating reference. Another is to fit a DC isolating device in the output, such as a magnetic amplifier.

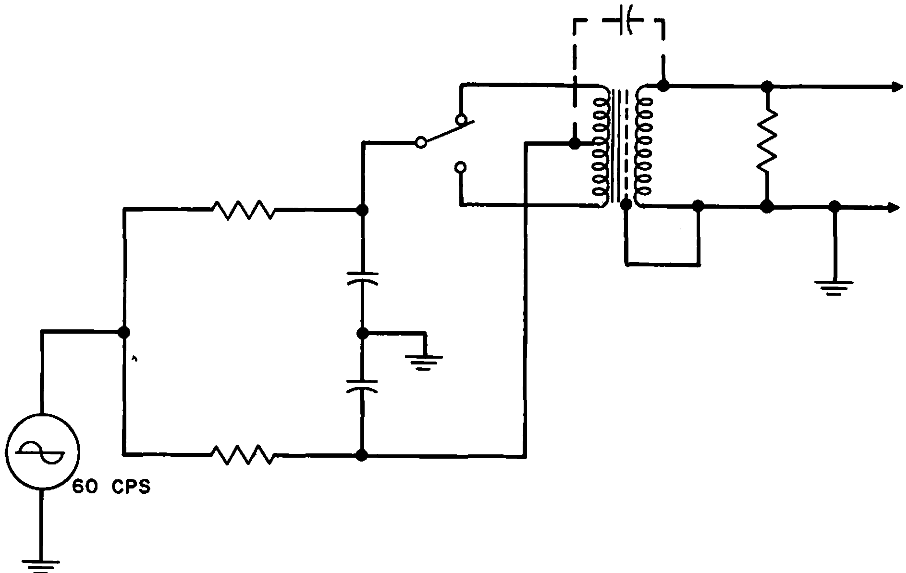


FIGURE 10. *Transformer coupling permits input at any DC level and greatly reduces common mode effect.*

Transformer coupling is also advantageous in that it enables the source to be matched to the amplifier. If the source resistance is a few kilohms, and the amplifier input impedance for example a megohm, then the transformer can have a step-up ratio of 1 to 10 or 20, and values of this order are used in self-balancing potentiometer recorders with thermionic valve amplifiers. This particular advantage is lost if the source and amplifier have about the same impedance, as is more often the case with transistor amplification.

Transformers suitable for very low level operation are rather special, and their design should not be undertaken lightly. A high permeability

nickel-iron core is used, and every winding (including part windings) must be wound in two equal parts on opposite legs so as to cancel out induced voltages. Even so, a high permeability external shield is almost always essential, and two or more shields with air spaces in between are not uncommon for more rigorous applications. One commercial design uses two 1/32 in. thick Mumetal shields and has a shielding efficiency of 60 *db*. To avoid loops outside the screen, the terminations are usually fly leads in twisted pairs or threes, not tabs or terminals. The effect of common mode pickup is directly proportional to the inter-winding capacity, and an efficient screen must be incorporated. Transformer inductance should be measured with small signals, since the initial permeability of the core may be 3 or 4 times less than the large signal value. In extreme cases it may be necessary to demagnetize the entire assembly to prevent microphonic effect, i.e. signals induced by the flexure of the whole structure. The literature contains several useful references on transformer design for this type of application, and there are several models on the market.

DC ISOLATION. Sometimes it is necessary to compare two signals which are at different levels relative to ground, and almost the only solution is an isolating circuit using a floating capacitor switched between one part of the circuit and the other. Figure 11, shows an arrangement which finds at least one commercial application in the measurement of pH, the input signal coming from two electrodes which are one or two volts away from ground, and requiring balancing by a feedback signal. Transformer coupling and a floating feedback are not practicable in this application, on account of the high resistances involved. For steady state analysis, the isolating circuit reduces to the simpler equivalent circuit shown in the same figure.

There is something to be said for operating this system at low frequencies. The transfer capacitor and its connected contacts inevitably have small capacities to ground and to the other contacts, and these are repeatedly charged to the potential of the input circuit and then discharged again, this happening once a cycle. The effect is to draw through the source resistances, currents which are proportional to frequency and which alter the potential across the filter capacitor, so causing an offset. This error is reduced proportionately by reducing the frequency, and the pH system mentioned previously operates at around 1 CPS, the chopper in this case being a British Post Office type 3000 relay with special contacts and very good insulation. Source resistances up to 1000 megohms are permissible, and accuracy is better than 1% with an input range of 0 to 100 mv.

THE OUTPUT FILTER. Whether an amplifier contains valves or transistors, a direct potential exists between its input terminals, and some form of AC coupling is essential to prevent small direct currents from reaching the chopper. Any form of DC leakage to the chopper is likely to give rise to an AC output, and leakages as small as 10^{-11} ampere may

cause trouble in sensitive modulators. Hence the coupling capacitor must be of a suitably high quality. If the current sensitivity of the system is high, or the chopping frequency low so that a large actual capacity is needed, it may be impossible to achieve the desired insulation resistance in a single capacitor, and in this case a two-stage CR output filter will solve the problem. The voltage across the first capacitor, due to leakage through the second, is so small that leakage through the first one can be neglected.

Precisely the same trouble can be caused by slow variations in the potential between the amplifier terminals, even if the capacitor is perfect. Heater or high voltage changes in valve amplifiers, or temperature-induced

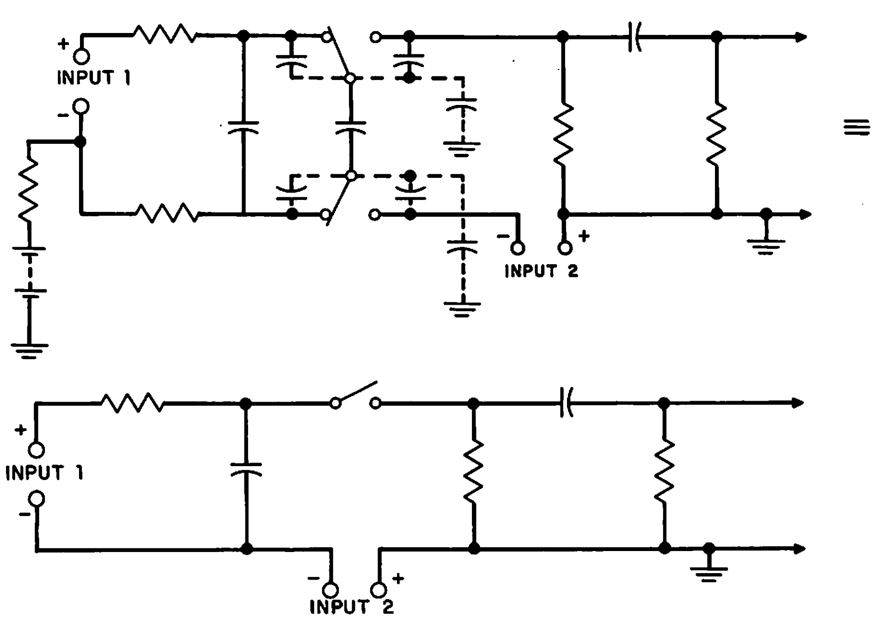


FIGURE 11. *DC isolating circuit is equivalent to grounded circuit in the steady state.*

changes in base to emitter voltage in transistor amplifiers may dispatch slowly varying currents to the chopper, causing in some cases disastrous temporary swings in the demodulated output. The solution is the same, namely to use a two or three stage high-pass output filter which transmits the chopped signal forward, but attenuates slow changes in the opposite direction. The arrangement is shown in Figure 12, and in the steady state is equivalent to a single resistance load whose value is equal to the filter resistances in parallel. Balanced input arrangements are considerably better than one-sided circuits, and symmetrical transformer-coupled arrangements are fairly immune.

Very occasionally, if the first amplifying stage is, for example, an electrometer valve having extremely low grid current, the coupling capacitor may be dispensed with altogether. Modulation is still useful in eliminating the slow DC drifts of such an amplifier.

BBM OR MBB? Single-pole double-throw circuits (and two-pole changeover circuits in the case of transformer coupling) can be either break before make or make before break. So far, BBM has been implied, but each system has its advantages.

Capacity-coupled BBM circuits are most efficient when a filter capacitor is connected across the input (Figure 13). The amplifier then sees a low

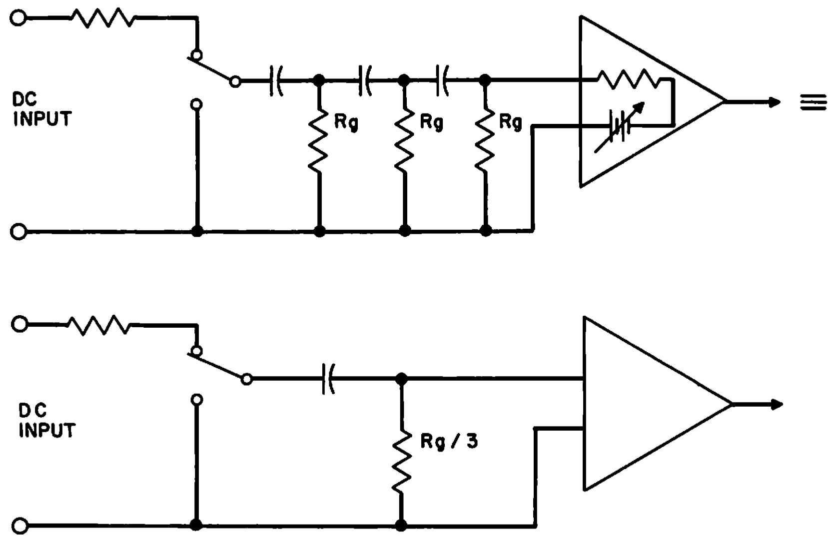


FIGURE 12. *High-pass output filter reduces effect of potential variations at amplifier input.*

AC impedance to ground at all times except during the transit time of the chopper, and this may permit spikes to appear in the output signal. The MBB setting avoids this trouble, provided that the chopper contacts remain perfect and the setting does not drift into BBM, but a stopper resistance must be inserted after the filter capacitor, and the efficiency is lower. If the overlap time of the contacts is short, the stopper can be quite small, and the amplifier sees a fairly low impedance at all times. It is better, when possible, to use the BBM arrangement, and eliminate the pickup currents, etc., which cause the spikes, since the circuit is then more likely to continue working in the unlikely event of partial contact failure. It is still necessary to ensure that the chopper does not drift into the MBB state if

a filter capacitor is present, unless a stopper is included as a precaution.

Transformer-coupled circuits must either be MBB or else contain a buffer capacity (which may be winding self-capacitance) to prevent inductive spikes and a great loss of efficiency.

STEADY-STATE EQUIVALENT CIRCUITS. There is a great variety of practical modulator circuits, based on permutations of the principles already discussed, and to analyze their steady-state performance it is first necessary to reduce each to its simplest form. One is then left with a reasonable number of basic equivalent circuits to consider in detail.

Equivalents have already been shown for the DC isolating circuit and the multi-stage output filter. A multiple input filter is equivalent to a single-stage filter with the same series resistance (Figure 13), and two inputs summed at one contact can be reduced to a single equivalent input (Figure 14).

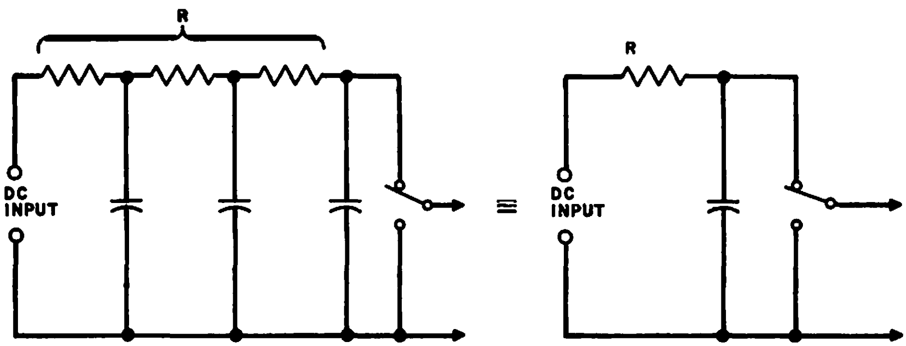


FIGURE 13. *Single and multi-stage input filters are equivalent in the steady state.*

These are fairly obvious points, but a little more care is needed when there are two separate inputs in series. Every modulator has a DC-carrying loop, and the two inputs are always in series in this loop, no matter how the AC signal is taken off. Each input can be represented by a voltage and a source resistance, and the effect of this resistance depends on whether the DC it carries is steady or pulsating, i.e., on whether or not there is a filter on that input. If the current is pulsating, the effect of the resistance depends not only on the mean value of the current, but also on the mark-space ratio of the current pulses. Thus a few rules can be deduced governing how circuits can be manipulated in deriving equivalents. But first we need two definitions.

VOLTAGE GAIN. The voltage gain of a modulator circuit is defined as the peak-to-peak output voltage divided by the DC input voltage.

INPUT RESISTANCE. This is defined as the (steady) input voltage divided by the mean input current.

RULES.

1. A given input voltage, injected in any part of the DC loop, always produces the same AC output. (Otherwise two equal and opposite signals would not give zero output.) In other words, the circuit gain may be referred to any input position and will always be the same for a particular circuit.
2. No matter where a given input voltage is injected, the mean DC circulated in the DC loop is always the same for a particular circuit. In other words the input resistance is the same at all points in series with the DC loop.
3. If two separate inputs are both filtered so that each source resistance carries steady DC, they can be considered as a single filtered input (Figure 15).

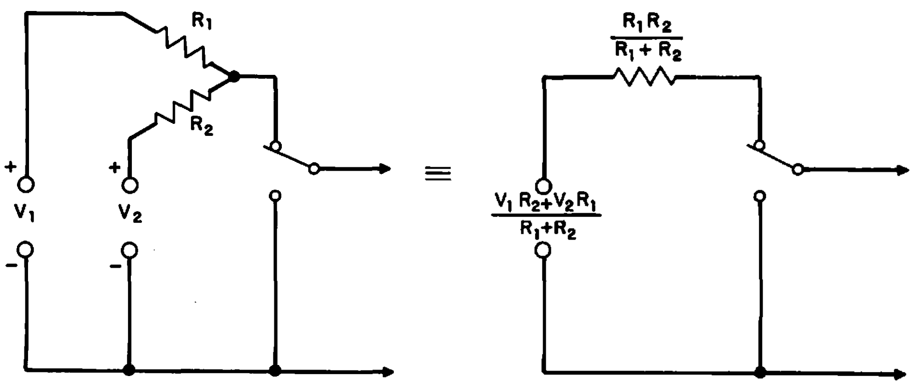


FIGURE 14. *Inputs summed on one contact reduce to a single input.*

4. If one input is filtered and the other is not, the circuit cannot be rearranged in this manner unless (as quite often occurs) one of the source resistances is negligible.
5. If neither input is filtered, the circuit cannot be rearranged unless the current pulses through the two source resistances have the same mark-space ratio. In general terms they do not, although in specific circuits they often may.

We can now list the basic capacity-coupled circuits which are needed to describe the steady-state operation of a very wide range of practical circuits. It turns out that there are six. There are three common switching arrangements:

1. On-off.
2. SPDT, BBM.
3. SPDT, MBB.

Both SPDT circuits have the moving contact connected to the output. One can envisage a two-pole changeover circuit as well, but it would seldom

be practicable on account of grounding problems. Each of these three arrangements can be used in circuits with:

1. Two unfiltered inputs.
2. One filtered and one unfiltered input.

If both inputs are filtered, they can be lumped together as one. These six basic circuits are analyzed in the next section. In each case it will be assumed that there are two inputs. In the on-off and SPDT, BBM circuits, each input will be assumed to have finite source resistance, the resulting expressions thus being as general as possible. It is easy to simplify them for the case where one source resistance vanishes, as often occurs in practice. In the SPDT, MBB case, one of the source resistances will perforce be assumed zero, since otherwise the arithmetic and the resulting expressions become rather heavy.

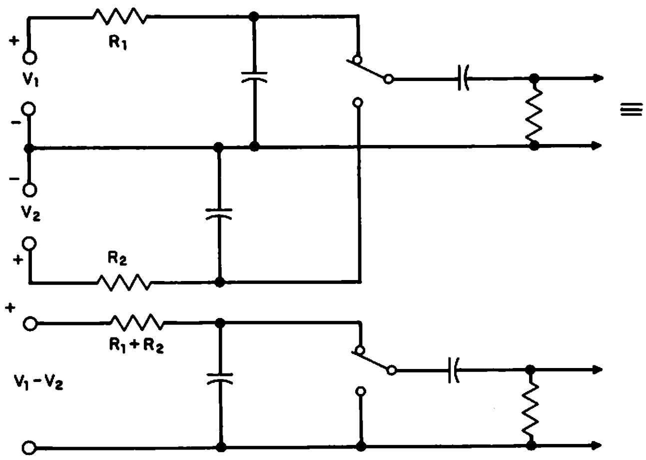


FIGURE 15. *Two filtered inputs can be regarded as a single filtered input.*

We can also list the basic transformer-coupled circuits. On-off versions are both inefficient and subject to pickup; if one takes the trouble to build in a transformer, it is worthwhile using a balanced arrangement. The possible switching arrangements are thus:

1. SPDT, BBM.
2. Two-pole changeover, BBM.
3. SPDT, MBB.
4. Two-pole changeover, MBB.

It is not feasible to feed a transformer from a high resistance source, so filtered inputs are assumed. If a feedback input is to be injected after the filter (a common arrangement), its source resistance will be assumed negligible (as it almost invariably is). Transformer-coupled circuits will be dealt with later.

SECTION III

ANALYSIS OF CAPACITY-COUPLED MODULATORS

Expressions for gain and input resistance of the six basic circuits are listed in Tables 1, 2 and 3, together with the value of stopper resistance (where one exists) giving greatest gain, the optimized gain expression, and the corresponding input resistance. Terminology is given on figures in the tables, x and y being contact closure times, and z, the common time expressed as fractions of a cycle. The method of analysis is similar for each circuit, and the following example is typical.

ON-OFF MODULATOR WITHOUT FILTER. Consider the on-off modulator with two unfiltered inputs, and for simplicity assume that input voltage No. 2 is zero (Figure 16). Both source resistances are retained, and there is no loss of generality. If we find the gain and input resistance for input No. 1, the results will apply also to input No. 2.

The coupling capacitor is assumed sufficiently large to pass on square waves without appreciable droop. It therefore assumes a steady potential, V_2, after a few initial cycles, and can be regarded as a battery. The circuit exists in one or other of two circuit states (Figure 16) during which steady current i_1, i_2, and i_3 flow through the resistors, being given by the following expressions:

$$V_1 = i_2(R_1 + R_2) + i_3 R_1$$
$$V_2 = i_2 R_2 - i_3 R_3$$
$$V_2 = i_1(R_2 + R_3)$$

V_1 being the input voltage.

Since no charge accumulates on the coupling capacitor, we have a fourth expression:

$$xi_3 = (1 - x)i_1$$

which, with the other three, permits the currents to be expressed in terms of V_1 and the resistances. The gain of the circuit is the peak-to-peak output voltage divided by the input voltage, i.e.,

$$\text{Gain, } A = \frac{(i_1 + i_3)R_3}{V_1}$$

which, with the values of i_1 and i_3 inserted, becomes:

$$A = \frac{1}{\left[1 + \frac{R_1}{R_2} + \frac{xR_2}{R_3} + \frac{R_1}{R_3}\right]}$$

The input resistance is the input voltage V_1, divided by the mean input current, i.e.,

$$\text{Input Resistance, } R_{in} = \frac{V_1}{(i_2 + i_3)x}$$

which becomes:

$$R_{in} = \frac{R_1}{x} + R_2 + \left(\frac{1-x}{x}\right)\left(\frac{R_2 R_3}{R_2 + R_3}\right)$$

These expressions hold for both inputs. For greatest gain, R_1 should

be as low as possible, and R_2 should have the optimum value, R_2 (opt) $= \sqrt{R_1 R_3 / x}$, obtained by differentiating the expression for gain with respect to R_2.

Inspection of the expression for gain shows that if there is only one source having appreciable resistance, it is better to make this No. 2 input, in which case R_1 can be made zero and R_2, which should then be as low as possible, is the source resistance in question. In other words, the shunt-switching arrangement is better than the series-switching arrangement. If, for some reason, the only source having appreciable resistance has to be made No. 1 input, then R_2 is a circuit resistor and should be given its optimum value. If there are two sources, each with appreciable resistance, then R_1 should be the lower of the two if the only consideration is obtaining the greatest possible gain. In practice, stability considerations will probably decide which input is which.

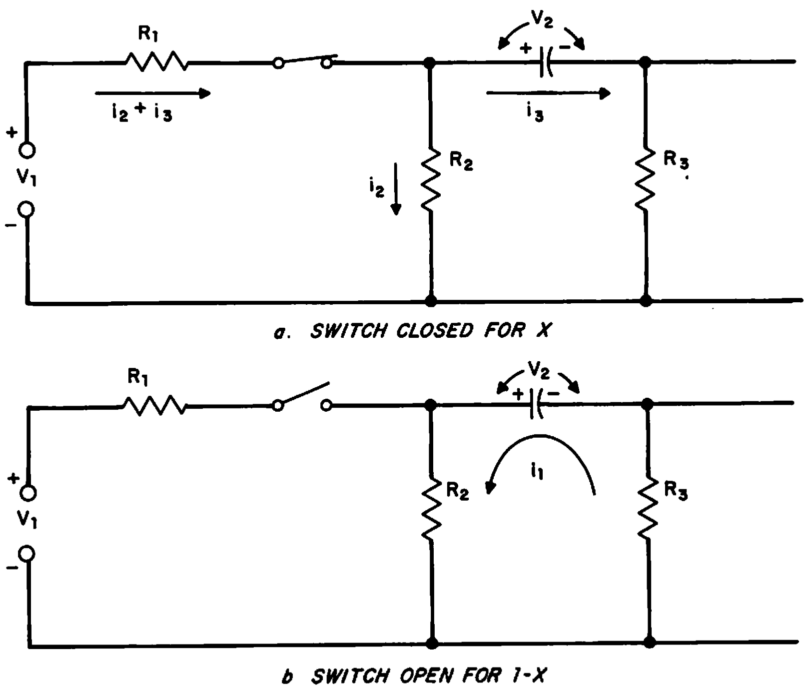


FIGURE 16. *On-off modulator. Circuit states.*

ON-OFF MODULATOR WITH FILTER. It is sufficient to consider the circuit with one input filtered and the other not filtered. Should both be filtered, they can be lumped together and considered as a single input. It matters not which input we consider, so let it be input No. 1.

Assume then, that a filter capacitor is connected across input No. 1 after the source resistance R_1, which becomes part of the filter. There may

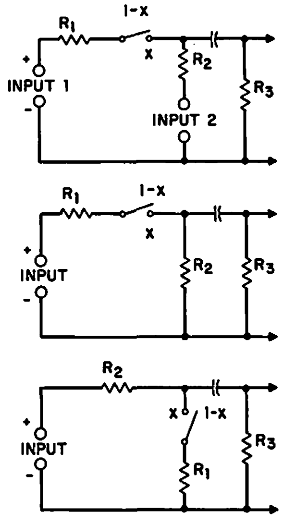
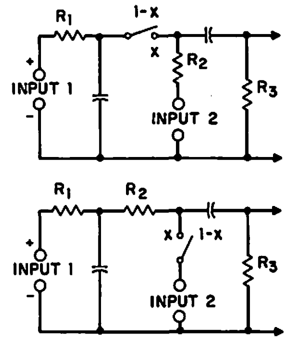
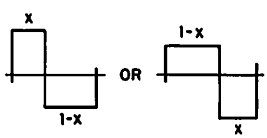
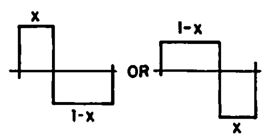
	WITHOUT FILTER	WITH FILTER
CIRCUIT AND SOME EQUIVALENTS		
WAVEFORM		
VOLTAGE GAIN	$\frac{1}{1+\frac{R_1}{R_2}+x\frac{R_2}{R_3}+\frac{R_1}{R_3}}$	$\frac{1}{1+x\left[\frac{R_1}{R_2}+\frac{R_2}{R_3}+\frac{R_1}{R_3}\right]}$
INPUT RESISTANCE	$\frac{R_1}{x}+R_2+\left(\frac{1-x}{x}\right)\left(\frac{R_2 R_3}{R_2+R_3}\right)$ (MEAN VALUE)	$R_1+R_2+\left(\frac{1-x}{x}\right)\left(\frac{R_2 R_3}{R_2+R_3}\right)$
R_2 (opt) FOR MAXIMUM GAIN	$\sqrt{\frac{R_1 R_3}{x}}$	$\sqrt{R_1 R_3}$
MAXIMUM GAIN	$\frac{1}{1+2\sqrt{\frac{xR_1}{R_3}}+\frac{R_1}{R_3}}$	$\frac{1}{1+2x\sqrt{\frac{R_1}{R_3}}+x\frac{R_1}{R_3}}$

TABLE 1. *Properties of on-off circuits.*

in fact, be separate source and filter resistances, but it makes no difference. The circuit is depicted in Table 1. The gain and input resistance from this capacitor onwards are obtained by writing $R_1 = 0$ in the expressions already obtained.

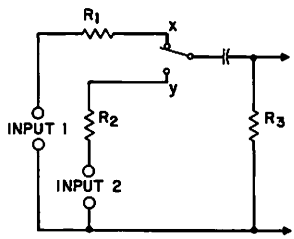
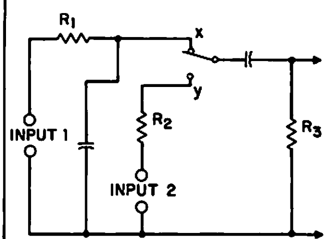
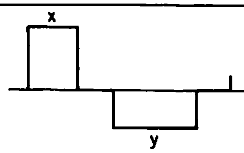
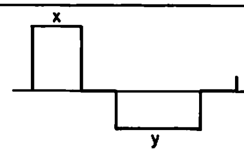
	WITHOUT FILTER	WITH FILTER
CIRCUIT		
WAVEFORM		
VOLTAGE GAIN	$\dfrac{1}{1+\left(\frac{xy}{x+y}\right)\left[\frac{1}{x}\frac{R_1}{R_3}+\frac{1}{y}\frac{R_2}{R_3}\right]}$	$\dfrac{1}{1+\left(\frac{xy}{x+y}\right)\left[\frac{R_1}{R_3}+\frac{1}{y}\frac{R_2}{R_3}\right]}$
INPUT RESISTANCE	$\frac{R_1}{x}+\frac{R_2}{y}+\left(\frac{x+y}{xy}\right)R_3$	$R_1+\frac{R_2}{y}+\left(\frac{x+y}{xy}\right)R_3$

TABLE 2. *Properties of SPDT, BBM circuits.*

The gain from input No. 1 is therefore:

$$\text{Gain} = A_{(R_1=0)} \times \frac{R_{in\,(R_1=0)}}{R_1 + R_{in\,(R_1=0)}} = \frac{1}{1 + x\left[\frac{R_1}{R_2} + \frac{R_2}{R_3} + \frac{R_1}{R_3}\right]}$$

And the input resistance from the same point is:

$$\text{Input resistance} = R_1 + R_{in(R_1=0)} = R_1 + R_2 + \left(\frac{1-x}{x}\right)\left(\frac{R_2 R_3}{R_2 + R_3}\right)$$

Again these expressions apply for both inputs. Again R_1 should be as low as possible, and preferably zero, while R_2 should have an optimum value which in this case is R_2 (opt) $= \sqrt{R_1R_3}$. The circuit can be arranged

in two ways, as shown in Table 1, by reversing the positions of R_2 and the switch. These are clearly identical, since the alternating voltage across R_2 must be the same as that across the switch.

COMPARISON OF CIRCUIT EFFICIENCIES. As we have already seen, there are many factors which govern the choice of circuit type, and

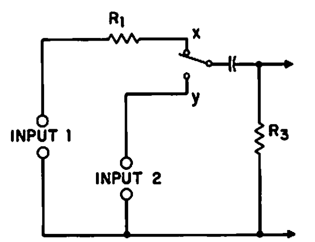
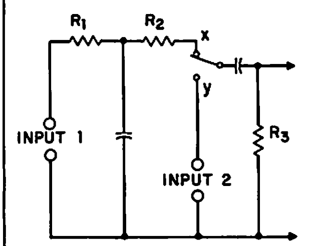
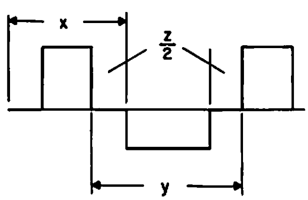
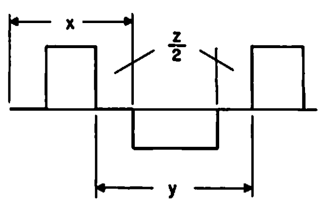
	WITHOUT FILTER	WITH FILTER
CIRCUIT		
WAVEFORM		
VOLTAGE GAIN	$\dfrac{1}{1+y\dfrac{R_1}{R_3}}$	$\dfrac{1}{1+z\dfrac{R_1}{R_2}+y\dfrac{R_2}{R_3}+xy\dfrac{R_1}{R_2}}$
INPUT RESISTANCE	$\dfrac{R_3+yR_1}{z\dfrac{R_3}{R_1}+xy}$	$R_1+\dfrac{R_3+yR_2}{z\dfrac{R_3}{R_2}+xy}$
R_2 (opt) FOR MAXIMUM GAIN		$\sqrt{\dfrac{z}{y}R_1R_3}$
MAXIMUM GAIN		$\dfrac{1}{1+2\sqrt{yz\dfrac{R_1}{R_3}}+xy\dfrac{R_1}{R_3}}$

TABLE 3. *Properties of SPDT, MBB circuits.*

efficiency is only one of them. The main value of the expressions listed in Tables 1 to 3 is therefore in assisting the choice of component values, and assessing the performance of an already chosen configuration. Nevertheless it is interesting to compare the circuits on the score of efficiency.

NOTES: $h = R_S/R_3$ R_2 = OPTIMUM WHEN PRESENT

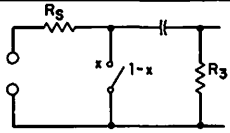
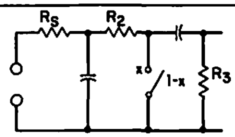
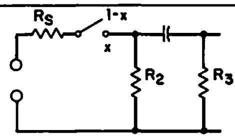
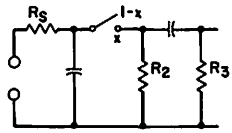
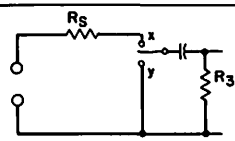
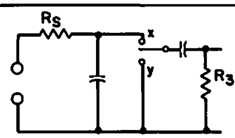
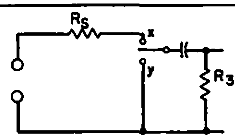
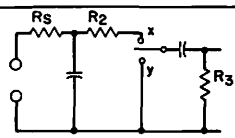
DESCRIPTION	CIRCUIT	VOLTAGE GAIN (h USUALLY < 1)	CURRENT/VOLTAGE GAIN (h USUALLY > 1)
ON-OFF SHUNT		$\frac{1}{1+xh}$	$\frac{1}{R_S}\cdot\frac{1}{\frac{1}{h}+x}$
ON-OFF SHUNT + FILTER		$\frac{1}{1+2x\sqrt{h}+xh}$	$\frac{1}{R_S}\cdot\frac{1}{\frac{1}{h}+2x\sqrt{\frac{1}{h}}+x}$
ON-OFF SERIES		$\frac{1}{1+2\sqrt{xh}+h}$	$\frac{1}{R_S}\cdot\frac{1}{\frac{1}{h}+2\sqrt{\frac{x}{h}}+1}$
ON-OFF SERIES + FILTER		$\frac{1}{1+2x\sqrt{h}+xh}$	$\frac{1}{R_S}\cdot\frac{1}{\frac{1}{h}+2x\sqrt{\frac{1}{h}}+x}$
SPDT, BBM		$\frac{1}{1+\left(\frac{y}{x+y}\right)h}$	$\frac{1}{R_S}\cdot\frac{1}{\frac{1}{h}+\frac{y}{x+y}}$
SPDT, BBM + FILTER		$\frac{1}{1+\left(\frac{xy}{x+y}\right)h}$	$\frac{1}{R_S}\cdot\frac{1}{\frac{1}{h}+\frac{xy}{x+y}}$
SPDT, MBB		$\frac{1}{1+yh}$	$\frac{1}{R_S}\cdot\frac{1}{\frac{1}{h}+y}$
SPDT, MBB + FILTER	 z = COMMON TIME PER CYCLE	$\frac{1}{1+2\sqrt{yzh}+xyh}$	$\frac{1}{R_S}\cdot\frac{1}{\frac{1}{h}+2\sqrt{\frac{yz}{h}}+xy}$

TABLE 4. *Gains of circuits having only one source resistance.*

To simplify matters, we shall assume that one of the two source resistances is negligible, as is almost always the case in practice. The circuits can then be compared on the basis of the gain which each provides for a given ratio h, of source to load resistance. (R_2, when present, is assumed to have its optimum value.) It makes no difference whether we compare voltage gains, or current/voltage gains, i.e., peak-to-peak load current/input voltage, and both are listed in Table 4.

The basic two-input on-off circuit here divides into two distinct arrangements according to which input position we consider, one having the switch in series with the input chosen, the other having it in shunt across the input. We thus have eight cases, though the two filtered on-off circuits, which look different at first sight, are in fact the same. Bearing in mind that x and y are less than unity, inspection of Table 4 shows:

1. The on-off circuits are in the following descending order of efficiency.
   a. On-off shunt.
   b. On-off shunt with filter } equal.
      On-off series with filter }
   c. On-off series.
2. The SPDT, BBM circuit is more efficient if its input is filtered. If $x = y = \frac{1}{2}$, it is better than any of the on-off circuits.
3. BBM is better than MBB, but there is not a lot of difference if the common time, z, is small.
4. Addition of a filter capacitor improves the efficiency in all cases where this capacitor is not intermittently shorted. It still does so in the SPDT, MBB circuit if the common time is short.

WAVEFORM SYMMETRY. All the on-off circuits provide more voltage gain if the switch closure time, x, is made small. This is not very significant, however, since h is usually small in cases where the criterion is voltage gain, and this accordingly usually approaches unity.

If the modulator works into a low resistance, however, such as a transistor amplifier, h may well be high, and we are concerned with how much current is delivered to the load, not with the voltage developed across it. Expressions for current/voltage gain are given in the last column of Table 4, and it can be seen that, if h is large, it may be worthwhile making x small in the on-off circuits (and y small in the others). In the limit, if the load is negligible, the current delivered to it is inversely proportional to x (or y). For this reason, asymmetrical switching is sometimes used when the load is low and current output is the criterion. There is a disadvantage, however, in that the demodulator must be able to supply more current during the shortened part of the waveform if it is to provide the same DC output.

SECTION IV

ANALYSIS OF MBB TRANSFORMER-COUPLED MODULATORS

In this section we shall consider two very similar transformer-coupled circuits, both having contact actions which ensure that the transformer primary is never open-circuited or presented with a high impedance. This eliminates the spikes which otherwise would occur unless the transformer were buffered by capacity. Both circuit arrangements are symmetrical ones, and in both cases we assume symmetrical switching as well, since asymmetry would give no advantage.

One of the circuits uses a center-tapped primary together with an SPDT, MBB chopper, while the other has a single primary and a two-pole changeover chopper; the latter may in practice be two SPDT, MBB units operated synchronously. The SPDT arrangement is the more common, since the chopper is simpler; additional complexity in the transformer is of less significance. The changeover circuit has the moderate advantage of eliminating the effect of any thermal e.m.f.'s generated within the transformer, but these can be kept small by suitable design.

From the analytical point of view, the only difference between the two arrangements is that the changeover circuit may contain a brief period (t_4) during each changeover action, when the source is not shorted out. This may occur if the two choppers are not well enough synchronized, and does not matter very much in practice, since bad synchronism never by itself open circuits the primary. In the SPDT circuit, the source is either connected to one of the half primaries or else shorted out, and t_4 does not exist.

The changeover circuit is therefore slightly more general, and is used as the basis for an analysis in which no major approximations are made. This gives rise to rather unwieldy expressions which are simplified according to suitable approximations, one of which is that $t_4 = 0$.

TERMINOLOGY.

$B = R_e t_2 / L$.

i = Instantaneous magnetizing current at time t, amperes.

$i_a = i + i_b$, amperes.

i_b = Instantaneous current in load, referred to primary, amperes.

$-i_o$ = Value of i at $t = 0$, amperes.

$+i_o$ = Value of i at $t = t_2$, amperes.

I = Mean current drawn from C_1, amperes.

k = Secondary/primary turns ratio, changeover circuit,
= Secondary/half primary turns ratio, SPDT circuit.

L = Inductance of primary winding, changeover circuit, henrys,
= Inductance of half primary winding, SPDT circuit, henrys.

TERMINOLOGY (Cont'd)

Q_2 = Net charge drawn from C_1 during t_2, coulombs.
R_1 = Filter resistance, ohms.
R_2 = Stopper resistance, ohms.
R_3 = Load resistance, ohms.
R'_3 = R_3/k^2, ohms.
R_s = Secondary winding resistance, ohms.
R'_s = R_s/k^2, ohms.
R_p = Resistance of primary winding, changeover circuit, ohms.
$4R_p$ = Resistance of primary winding, SPDT circuit, ohms.
R_a = $R_2 + R_p$, ohms.
R_b = $R'_3 + R'_s$, ohms.
R_o = Input resistance, measured at C_1, ohms.
R_e = $R_a R_b/(R_a + R_b)$, ohms.
t_1 = Cycle time, second.
$2t_2$ = Pulse time per cycle, second.
$4t_3$ = Common time per cycle, second.
$2t_4$ = Open time per cycle, changeover circuit, second,
= Zero, SPDT circuit.
V = Direct input voltage, volts.
V_1 = Voltage across C_1, volts.
V_3 = Twice mean output voltage during t_2, volts.

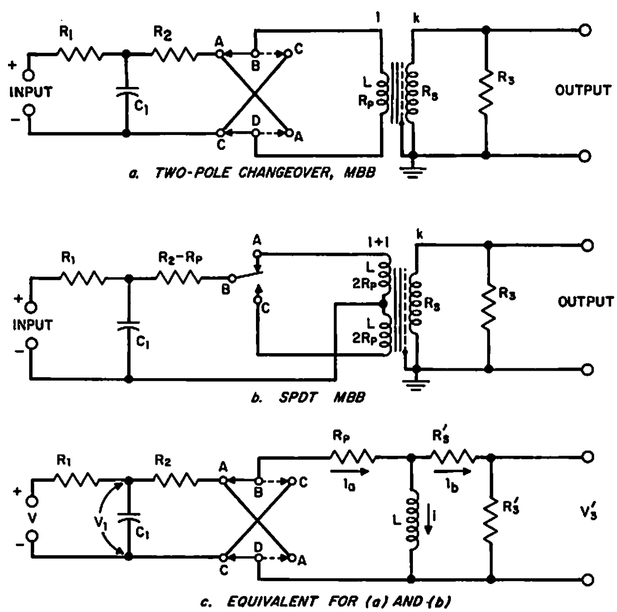


FIGURE 17. *MBB transformer — coupled modulators.*

THE GENERAL CASE. The two practical circuits and their single equivalent circuit are shown in Figure 17. Both need an input filter so that the transformer is fed from a low-impedance source and the waveform is as square as possible. Both need a stopper resistance to control the current drain from the filter capacitor during switchover. The changeover switch is assumed to be made up of two MBB, SPDT choppers, each having a common time per cycle of $2t_3$, and slightly out of synchronism. This is equivalent to a common time per cycle of $4t_3$ in the SPDT circuit. When all three contacts of the SPDT circuit are joined, the shorted primary

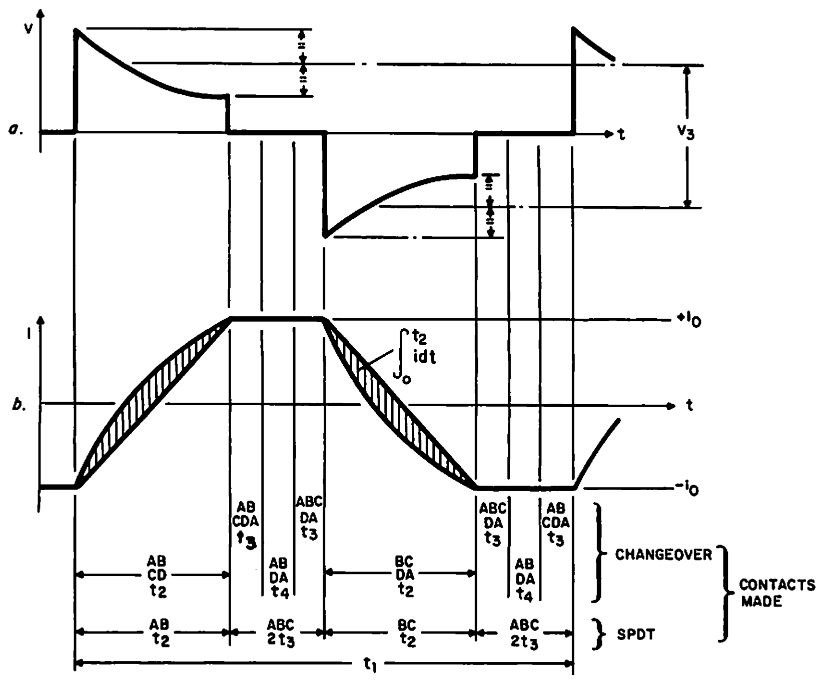


FIGURE 18. *Waveforms of output voltage and magnetizing current.*

looks to the filter like a resistance R_p, since equal and opposite currents flow in the two half primaries, and induce no net flux or back e.m.f. The stopper resistor is therefore labelled $R_2 - R_p$, so that the single equivalent circuit is true for both cases. Transformer leakage reactance and self capacitance are neglected, but losses can be accounted for if necessary by assuming the load resistance to include the transformer loss resistance.

The filter capacitance is assumed large, so that steady voltage exists across it when steady-state conditions have been reached after a number of initial cycles. Figure 18, shows the typical output voltage waveform in which both the droop and the common time have been exaggerated for

clarity. The magnetizing current waveform is also shown. This can be observed practically on a changeover circuit with the load removed. In the SPDT arrangement, the primaries carry the filter short-out current in addition, as shown in Figure 19.

Operation of the SPDT circuit is considered in rather more detail later, and shown to be essentially the same as that of the equivalent circuit. Consider therefore the equivalent circuit, Figure 17c. At time $t = 0$, the filter capacitor is connected via R_2 to the transformer, and currents flow in the circuit, whose instantaneous values are i_a, i and i_b. After a

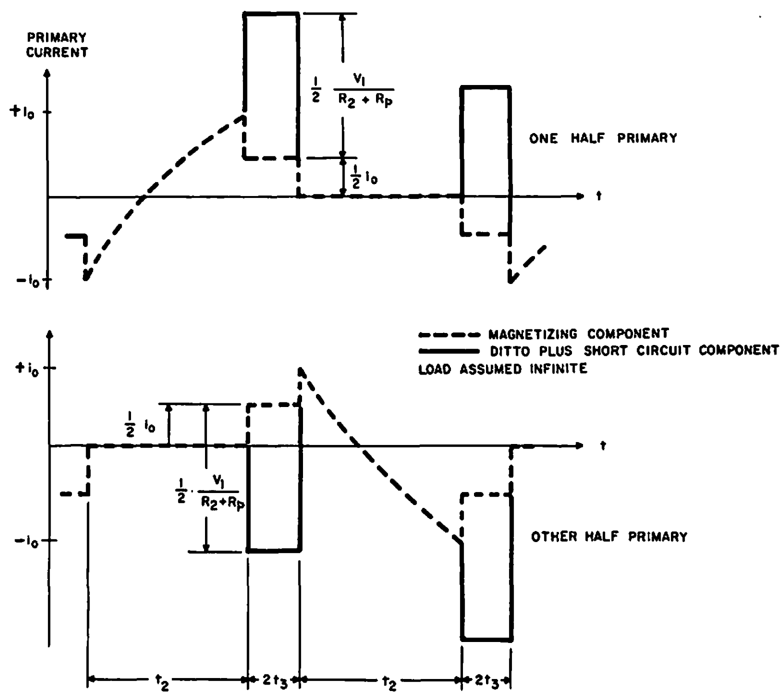


FIGURE 19. *Primary current waveforms. SPDT case.*

time t_2, the current in L has risen to a value $+ i_o$ and the transformer is short-circuited for a time $t_4 + 2t_3$. During this short-circuit period, the flux in the transformer core is maintained by i_o which remains sensibly constant since the primary self time constant, L/R_p is relatively long.

The next half cycle therefore commences with a current $+ i_o$ still flowing through L and tending to charge the capacitor, now connected the other way round, back to its original potential. The current falls during t_2 to a final value which, by symmetry, must be $-i_o$, and there is then another short circuit period, after which the cycle begins again. Because of the

symmetry, it is sufficient to analyze a single half cycle. This gives expressions for the currents, the mean drain from the capacitor, the input resistance and the gain.

Let
$$R_a = R_2 + R_p$$
$$R_b = R'_s + R'_3$$
$$R_e = \frac{R_a R_b}{R_a + R_b}$$

Then the basic circuit equations for the period t_2 are:

$$V_1 = R_a i_a + L\,\delta i/\delta t$$
$$L\,\delta i/\delta t = R_b i_b$$
$$i_a = i + i_b$$

Eliminating i_a and i_b from the above, integrating, and putting $i = -i_o$ when $t = 0$, and $i = +i_o$ when $t = t_2$, gives:

$$i = \frac{V_1}{R_a}\left(1 - \frac{2\epsilon^{-R_et/L}}{1 + \epsilon^{-R_et/L}}\right)$$

This enables us to find the charge drawn from C_1 during t_2, which is:

$$Q_2 = \int_0^{t_2} (i + i_b)\,\delta t = \int_0^{t_2} i\delta t + \frac{L}{R_b}\Big[\,i\,\Big]_0^{t_2}$$
$$= \frac{V_1}{R_a}\left[t_2 - 2\,\frac{L}{R_a}\left\{\frac{1 - \epsilon^{-R_et_2/L}}{1 + \epsilon^{-R_et_2/L}}\right\}\right]$$
$$= \frac{V_1}{R_a}\left[t_2 - 2\,\frac{L}{R_a}\,\tanh\,\frac{1}{2}\,\frac{R_et_2}{L}\right]$$

For convenience, the *tanh* term is now replaced by the first two terms of its expansion as a power series. Provided that R_et_2/L is not greater than unity, the consequent error in Q_2 is below 1%, and we are not interested in greater values, since they imply a peaky waveform whose droop in fact exceeds 63%. Thus:

$$Q_2 = \frac{V_1}{R_a}\left[\frac{R_at_2}{R_a + R_b} + \frac{1}{12}\,\frac{R_e^3\,t_2^3}{R_a L^2}\right]$$

Now the charge drawn from C_1 during a complete half cycle is:

$$Q_2 + 2t_3V_1/R_2$$

the second term representing the short-circuiting of C_1 through R_2 during the switchover action, and the mean current drain from C_1 is therefore:

$$I = \frac{2}{t_1}\left[Q_2 + 2t_3\,V_1/R_2\right]$$

The input resistance of the circuit, measured at C_1 is thus:

$$R_o = \frac{V_1}{I} = \frac{R_a + R_b}{\frac{2}{t_1}\left[2t_3\left(\frac{R_a + R_b}{R_2}\right) + t_2 + \frac{1}{12} R_e^2 \frac{R_b t_2^3}{R_a L^2} \right]}$$

The input resistance, measured at the input terminals of the circuit, is of course obtained by adding R_1 to the above value of R_o.

We can now find the gain. The voltage across L at any time during t_2 is:

$$L\frac{\delta i}{\delta t} = 2V_1 \frac{R_e}{R_a}\left[\frac{\epsilon^{-R_e t/L}}{1 + \epsilon^{-R_e t_2/L}}\right]$$

and the sum of its values at $t = 0$ and $t = t_2$ is:

$$2V_1 R_e/R_a$$

Multiplying this by R'_3/R_b, we obtain the output across R'_3, namely:

$$2V_1 R_e R'_3/R_aR_b$$

and the gain, from the filter capacitor onwards, is:

$$2k R'_3/(R_a + R_b)$$

where k is the turns ratio of the transformer. The gain from the input terminals is obtained by multiplying this expression by $R_o/(R_1 + R_o)$ and is:

$$\text{Gain} = \frac{2k}{\frac{R_a + R_b}{R'_3} + \frac{2R_1}{t_1 R'_3}\left[2t_3\left(\frac{R_a + R_b}{R_2}\right) + t_2 + \frac{1}{12}\frac{R_b}{R_a}\frac{R_e^2}{L^2} t_2^3\right]}$$

This has its maximum value, with R_2 varying when:

$$2t_3\left(\frac{R_b + R_p}{R_2^2}\right) = \frac{1}{2}\frac{t_1}{R_1} + \frac{1}{12}\frac{t_2^3}{L^2}(R_b - R_a)\left(\frac{R_b}{R_a + R_b}\right)^3$$

APPROXIMATIONS. The expression for gain just developed is accurate to 1% provided only that C_1 is very large and $R_e t_2/L$ is not greater than unity. Beautiful though it may be in its generality, it is not of very much use as it stands. But for those who want to make their own deductions, or who have an extraordinarily general transformer,it may be useful.

It seems that there are several ways of deriving simpler approximate expressions, and we shall start by making the following assumptions:

1. Neglect the winding resistances. This gives little error in most practical cases, unless one is using a transformer with far more inductance than necessary, and hence too much winding resistance. Having designed a circuit on this assumption, it is simple to check that, in fact, $R_p \ll R'_3$, and $R_s \ll R_3$.

2. Put $t_4 = 0$, so that in consequence $2t_2 + 4t_3 = t_1$. This is always true of the SPDT circuit, and the changeover circuit would always be set up so that it were true initially.

3. Write R'_3 as R_3/k^2 so that k appears explicitly in the expression.

4. Write $R_e t_2/L$ as B, a quantity which we have already assumed to be less than or equal to unity.

Assuming the foregoing, and no more, we can simplify the expression for gain to:

$$\text{Gain} = \frac{2}{k\left[\frac{R_2}{R_3} + \frac{R_1}{R_3}\right] + \frac{1}{k}\left[1 + 4\frac{t_3}{t_1}\frac{R_1}{R_2}\right] + \frac{1}{6k}\frac{R_1}{R_2}\frac{t_2}{t_1}B^2}$$

To simplify matters still further, and because L can often be made large, we next assume L to be infinite. Just how large L has to be to make this legitimate can be deduced very easily, but the resulting expression is not very informative. Thus:

$$\text{Gain } (L \text{ infinite}) = \frac{2}{k\left[\frac{R_2}{R_3} + \frac{R_1}{R_3}\right] + \frac{1}{k}\left[1 + 4\frac{t_3}{t_1}\frac{R_1}{R_2}\right]}$$

There is clearly a value of R_2 which makes this a maximum. Since R_2 is merely a circuit resistor, we are interested in this value which is:

$$R_2 \text{ (opt)} = \sqrt{4\frac{t_3}{t_1}\frac{R_1R_3}{k^2}} \quad \text{(for any value of } k\text{)}$$

Similarly, there is a value of transformer ratio, k, which makes the gain a maximum. If one is designing a circuit, this too is under control, and we are interested in it.

$$k \text{ (opt)} = \sqrt{\left(\frac{R_3}{R_1 + R_2}\right)\left(1 + 4\frac{t_3}{t_1}\frac{R_1}{R_2}\right)} \quad \text{(for any value of } R_2\text{)}$$

If R_2 and k are both optimized at the same time, we get:

$$R_2 \text{ (opt opt)} = R_1\sqrt{4\frac{t_3}{t_1}}$$

$$k \text{ (opt opt)} = \sqrt{\frac{R_3}{R_1}}$$

The last expression is the same as obtains for correct matching in sine wave circuits. Putting these simultaneous optimum values into the expression for gain (L not infinite) we obtain the optimum gain:

Gain (opt R_2, opt k) $=$

$$\left(\frac{1}{1+\sqrt{4\dfrac{t_3}{t_1}}}\right)\sqrt{\frac{R_3}{R_1}}\left[\frac{1}{1+\dfrac{1}{12}\left(\dfrac{1}{4\dfrac{t_3}{t_1}+\sqrt{4\dfrac{t_3}{t_1}}}\right)\dfrac{t_2}{t_1}B^2}\right]$$

$$= \frac{1}{1+\sqrt{4\dfrac{t_3}{t_1}}}\sqrt{\frac{R_3}{R_1}} \text{ approximately,}$$

$$\text{provided that } B \leqslant 1 \text{ and } \sqrt{4\frac{t_3}{t_1}} \gg \frac{1}{24}$$

The first condition has been assumed all along, and is necessary in any case if the waveform is to be a useful shape. The second requires a factor of inequality of e.g., 10 for 10% accuracy in the approximate optimum gain expression, or of 100 for 1% accuracy and so on. $4t_3/t_1$ is the fractional common time of the contacts, and might typically be 1/10, giving a factor of inequality of about 7.5, and accuracy of at least 15% in the approximate expression. Most of the above expressions are listed in Table 5, together with the assumptions on which they are based.

SIGNIFICANT PRIMARY RESISTANCE. Sometimes the resistance of the transformer primary is not entirely negligible. If a circuit is designed on the basis of expressions (Table 5) which neglect winding resistance altogether, and a practical transformer has primary resistance R_p, which is significant in relation to R_3/k^2, the following rather more accurate expressions may be of use. The assumptions are:

L infinite.

$t_4 = 0$.

R_s negligible (in a step-up transformer it is nearly always much less than R_3).

$R_p \ll (t_1/4t_3)\, R_2$ (i.e. R_p of the same order as R_2 or less).

And the resulting expressions are:

$$\text{Gain} = \frac{2}{k\left[\dfrac{R_a}{R_3}+\dfrac{R_1}{R_3}\right]+\dfrac{1}{k}\left[1+4\dfrac{t_3}{t_1}\dfrac{R_1}{R_2}\right]}$$

$$R_2\text{ (opt)} = \sqrt{4\frac{t_3}{t_1}R_1\left(\frac{R_3}{k^2}+R_p\right)}$$

$$k\text{ (opt)} = \sqrt{\left(\frac{R_3}{R_1+R_a}\right)\left(1+4\frac{t_3}{t_1}\frac{R_1}{R_2}\right)}$$

LOW TRANSFORMER RATIO. If the transformer ratio k is much smaller than the optimum, then R_3/k^2 becomes large compared with the

ASSUMED IN ALL EXPRESSIONS:

$$\begin{cases} \text{WINDING RESISTANCES NEGLIGIBLE} \\ t_4 = 0 \quad \text{(INHERENT IN SPDT CIRCUIT)} \\ B = \dfrac{R_2R_3/k^2}{R_2+R_3/k^2}\,\dfrac{t_2}{L} \ll 1 \quad (\text{FOR DROOP} \ll 63\%) \end{cases}$$

PROPERTY	EXPRESSION	ADDITIONAL ASSUMPTIONS
GAIN	$\dfrac{2}{k\left[\frac{R_2}{R_3}+\frac{R_1}{R_3}\right]+\frac{1}{k}\left[1+4\frac{t_3}{t_1}\frac{R_1}{R_2}\right]+\frac{1}{6k}\frac{t_2}{t_1}\frac{R_1}{R_2}B^2}$	NONE
GAIN	$\dfrac{2}{k\left[\frac{R_2}{R_3}+\frac{R_1}{R_3}\right]+\frac{1}{k}\left[1+4\frac{t_3}{t_1}\frac{R_1}{R_2}\right]}$	L INFINITE
R_2 (opt) FOR MAXIMUM GAIN	$\sqrt{4\frac{t_3}{t_1}\frac{R_1R_3}{k^2}}$	L INFINITE
k (opt) FOR MAXIMUM GAIN	$\sqrt{\left(\frac{R_3}{R_1+R_2}\right)\left(1+4\frac{t_3}{t_1}\frac{R_1}{R_2}\right)}$	L INFINITE
R_2 (opt opt)	$R_1\sqrt{4\frac{t_3}{t_1}}$	R_2 AND k BOTH OPTIMUM L INFINITE
k (opt opt)	$\sqrt{\frac{R_3}{R_1}}$	
GAIN (opt opt)	$\left(\dfrac{1}{1+\sqrt{4\frac{t_3}{t_1}}}\right)\sqrt{\frac{R_3}{R_1}}$	R_2 AND k BOTH OPTIMUM $\sqrt{4\frac{t_3}{t_1}} > \frac{1}{24}$ NO ADDITIONAL RESTRICTION ON L
TERM DESCRIBING OUTPUT WAVEFORM	ϵ^{-Bt/t_2}	NONE

TABLE 5. *Properties of MBB transformer — coupled modulators.*

optimum for R_2 based on the assumption of infinite L. The condition for this is:

$$\frac{R_3}{k^2} \gg \sqrt{4\,\frac{t_3}{t_1}\,\frac{R_1R_3}{k^2}}$$

$$\text{i.e.,}\quad k \ll \sqrt{\frac{t_1}{4t_3}\,\frac{R_3}{R_1}}$$

and may occur even with optimum k if the common time is very small. We can then write R_2 for R_e in the gain expression (neglecting winding resistances but not L), and find that:

$$R_2\ (\text{opt}) = \sqrt{\frac{24\,t_3}{\dfrac{t_2{}^3}{L^2} + 6\,\dfrac{t_1k^2}{R_1R_3}}}$$

which may depend on L. If, in addition, k or L are so low that:

$$k^2 \ll \frac{1}{6}\,\frac{t_2{}^3}{t_1}\,\frac{R_1R_3}{L^2}$$

$$\text{then } R_2\ (\text{opt}) = \frac{L}{t_2}\sqrt{\frac{24\,t_3}{t_2}}$$

In this case, since R_2t_2/L must be less than or equal to unity, we have:

$$24\,\frac{t_3}{t_2} \leqslant 1$$

as a further condition, i.e., the common time must be less than about 8% of the cycle. These expressions may be of use in the restricted case of improper matching (k too low) or of unusually short common time together with low inductance.

NOTE ON THE SPDT CIRCUIT. The analysis has been made in terms of an equivalent circuit which is, in fact, a changeover arrangement. One can see that this is justified by studying the operation of the SPDT circuit a little more closely. During t_2, the current in half the primary rises from $-i_o$ to $+i_o$, as in the equivalent. Then all three contacts are joined, and no potential can appear across the two ends of the primary. Hence the core flux remains almost constant, being maintained by a current $i_o/2$ which circulates through the whole primary. Superimposed on this are two equal currents which flow in opposite directions through the two half primaries under the influence of the input voltage. These set up no flux and no back e.m.f. so, to the input, the shorted transformer looks like a resistance R_p. These currents represent a drain on the filter and are controlled by the stopper; without it, they would be very large. As soon as the first half of the primary is disconnected, these currents cease, and the current in the second half immediately takes up the value $+i_o$, falling during t_2 to $-i_o$ again. Figure 19, shows the waveforms of magnetizing and of total cur-

rent in both half primaries with infinite load. The above rather intuitive description neglects possible interaction between the two components of current, but can be shown to be correct by detailed analysis.

CAPACITANCE ACROSS THE TRANSFORMER. If the transformer primary (both halves in the SPDT case) is open-circuited even briefly while current is flowing, the flux collapses and large voltage spikes are set up unless some capacitance exists across the winding. Apart from causing a bad waveform, this effect wastes the stored magnetic energy and lowers the circuit efficiency. A small amount of capacitance is therefore desirable, even in MBB circuits, to guard against the possibility of contact bounce or similar minor faults. Sufficient capacitance for this purpose is usually inherent in the transformer.

BBM circuits, of course, may need more buffering than this, and the capacitance should be chosen so that the spike, which is really the first part of what would become a train of dying oscillations, just meets the next horizontal part of the waveform at the end of the switching period. However, these circuits need separate analysis.

Sometimes one sees MBB circuits with sufficient shunt capacitance added to resonate the transformer magnetizing inductance at the operating frequency. This appears to be of no value at all. The capacitance is charged up during each t_2 period and promptly shorted out again during the switching period, the net effect being to drain current unnecessarily from the input. Capacitance of this order may lower the input resistance considerably. It does not even give appreciable discrimination against pickup which has been chopped into twice the operating frequency.

DESIGN PROCEDURE. The following example outlines the sort of procedure to be adopted in designing an MBB transformer-coupled circuit. Suppose that a SPDT circuit is required to work at 50 CPS from a source resistance of 1000 ohms into a load of 100,000 ohms and deliver a waveform whose droop is about 25%.

The common time must be large enough to guarantee continued MBB action in spite of contact drift, and 10% is a reasonable figure, giving $t_3 = 0.5$ m.s. This determines t_2. Winding resistances are neglected to start with, and R_2 and k are optimized assuming that L is infinite. (Table 5). Thus:

$t_1 = 20$ m.s.
$t_2 = 9$ m.s.
$t_3 = 0.5$ m.s.
$R_1 = 1000$ ohms.
$R_3 = 100{,}000$ ohms.
k (opt opt) $= 10$.
R_2 (opt opt) $= 1000\sqrt{2/20} = 320$ ohms approximately.
$R_3/k^2 = 1000$ ohms.
$R_e = 1000 \times 320/1320 = 240$ ohms approximately.

For 25% droop, B must be about 0.25, and hence L must be $4R_c t_2 = 8.6$ henrys. In practice a value of 10 henrys would be specified, measured with small signals applied. Now $\sqrt{4t_3/t_1} = 0.32$ which exceeds 1/24 by a factor of 7.5, so the optimum gain expression is accurate to 15% if $B = 1$. Since $B = 0.25$, it will be accurate to 1%, and the gain is $10/(1 + 0.32) = 7.5$. The secondary winding resistance of a transformer of this type would be negligible in relation to 100,000 ohms. The stopper resistor is given the value $R_2 - R_p$, and each half primary $(2 R_p)$ must therefore be less than 640 ohms, so that this is possible. Inspection of the circuit, Figure 17, shows that R_p should also be small in relation to R'_3, if expressions derived by neglecting R_p are to be true, so there may be a noticeable drop in gain if R_p exceeds about 100 ohms.

Should the transformer design require R_p to be, for example, 200 ohms, the more accurate expressions which allow for R_p may be used. Values of k and R_2 which simultaneously satisfy the expressions for optimum conditions are found by trial and error to be about 9 and 400 ohms respectively, and the corresponding gain is 7.0.

SECTION V
ANALYSIS OF BBM TRANSFORMER-COUPLED MODULATORS

BBM transformer-coupled modulators nearly always take one or the other of two practical forms, namely the two-pole changeover circuit (Figure 20a) or the SPDT circuit (Figure 20b), both of which are represented by the single equivalent circuit shown in Figure 20c.

Unless the source resistance is exceptionally low, it should be followed by a filter capacitor. This not only provides filtering and an output wave-

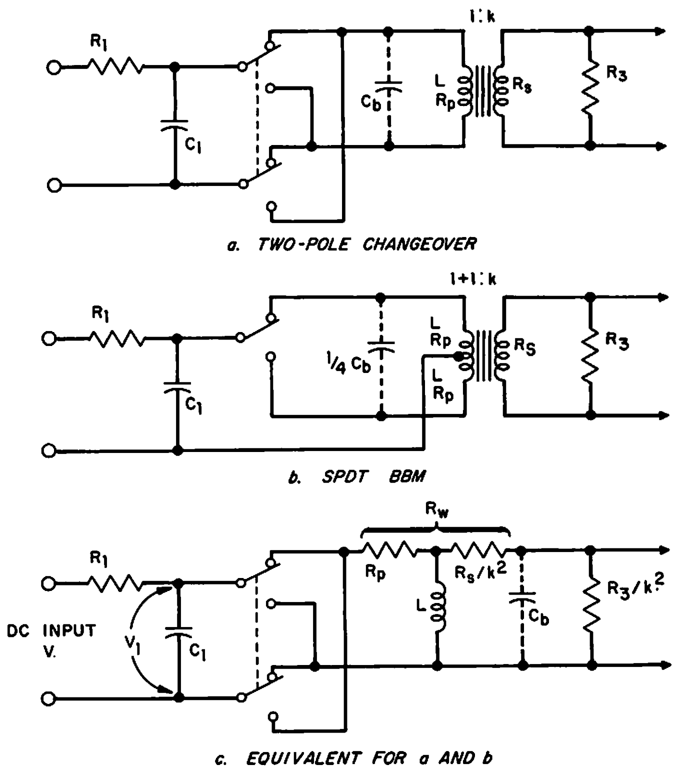


FIGURE 20. *BBM transformer — coupled modulators.*

form having an almost flat top, but also prevents losses which otherwise would be caused by the transformer magnetizing current flowing through the source resistance. The filter is never short circuited, and no stopper resistance is needed between filter and transformer. But a buffer capacitor, C_b, is frequently required to absorb the magnetizing current during the off time, and so control the voltage reversal which then occurs.

In what follows, transformer leakage reactance, self capacitance, and winding resistance are first neglected, as representing small departures from the ideal case. Terminology is shown in Figures 20 and 21, being largely the same as in the MBB case. The following differences are noteworthy:

$2t_2$ = Dwell time per cycle (= Pulse time per cycle), second.
$4t_3$ = Off time per cycle, second.
R_p = Resistance of primary winding, changeover circuit, ohms,
= Resistance of half primary, SPDT circuit, ohms.
R_w = $R_p + R_s/k^2$, ohms.
C_b = Buffer capacitance, referred to primary, mfd.

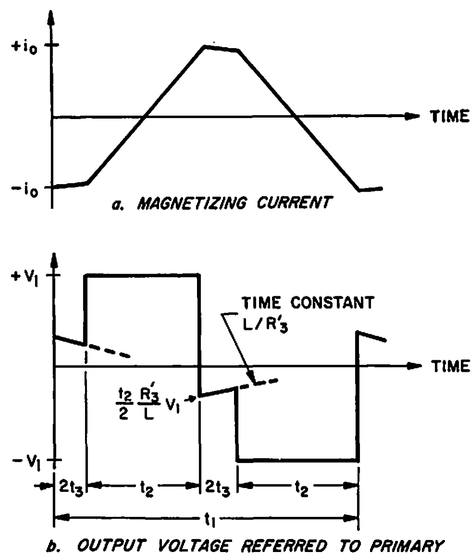


FIGURE 21. *Current and voltage waveforms — large L ($C_b = 0$).*

The mode of operation is rather different according to whether L is large or small, and the two cases (which adjoin) are taken separately.

OPERATION WITH LARGE L. Consider the equivalent circuit shown in Figure 20c. Suppose that L/R'_3 considerably exceeds $2t_3$ and that there is no buffer capacitor.

During one dwell period t_2, the magnetizing current in L rises to a value $+i_o$. The primary is then open circuited, and instantaneously a current i_o flows in the load R'_3. Because of our assumption about L, this current decreases only slightly during the off time, having nearly the same value at the start of the next dwell period, and flowing back into the filter capacitor C_1. Hence the magnetizing current swings between $+i_o$ and $-i_o$ approximately during each dwell period, as shown in Figure 21a. Since

L is large, the currents are small, and there is very little net DC drain from the filter capacitor which, however, prevents AC flowing through the source resistance and so causing I^2R losses.

During the dwell periods, the output voltage referred to the primary, is steady and equal to plus or minus the voltage V_1 on the filter capacitor. Winding resistance may in practice cause some, but as a rule not much, droop. The question is, what happens during the off periods?

The magnetizing current rises at a rate dictated entirely by L and V_1, and its value at the end of the dwell period is:

$$i_o = \frac{1}{2}\frac{t_2}{L}V_1$$

At the instant the primary is open circuited, this current flows in R'_3, giving a reverse voltage:

$$i_oR'_3 = \frac{1}{2}\frac{R'_3}{L}t_2V_1$$

which decays with time constant L/R'_3 until the next dwell period. For this reverse voltage to be less than or equal to V_1, as is usually required, we have the further condition that:

$$\frac{L}{R'_3} \geqslant \frac{t_2}{2} \qquad \text{(for no spike).}$$

A typical output voltage waveform is shown in Figure 21b. As L is increased, the intermediate steps sink towards the baseline.

This mode of operation is rather uneconomic in that it requires a fairly large value of inductance, but it does have the advantage that no buffering is needed. Although circuits with lower L are just as efficient (neglecting winding resistance) if they are correctly buffered, this condition is not maintained if the contact setting drifts. It is possible to round off the intermediate steps by adding capacity across L, but not possible to make the circuit oscillatory during the off time, since there is little energy associated with L. In practice there is usually some rounding off due to transformer self-capacitance.

GAIN WITH LARGE L. Assuming the off time to be a fairly small fraction of a cycle, the input resistance seen at C_1 is approximately $R'_3 = R_3/k^2$. This is a fair approximation, particularly if the height of the intermediate steps approaches V_1, since a voltage approximating $\pm V_1$ is then maintained across the load all the time. The energy to do this must come from somewhere, and is in fact supplied from C_1 which delivers a net current slightly in excess of V_1/R'_3 during the dwell period.

The voltage across C_1 is therefore:

$$V_1 = V\frac{R_3/k^2}{R_1 + R_3/k^2}$$

and the gain is:

$$\text{Gain} = 2\,k\frac{V_1}{V} = \frac{2\,R_3}{kR_1 + \frac{1}{k}R_3}$$

As k varies, this has a maximum value of:

$$\text{Gain (opt)} = \sqrt{R_3/R_1}$$
$$\text{given by } k \text{ (opt)} = \sqrt{R_3/R_1}$$

Thus correct matching requires the same transformer ratio as would be used in sine wave circuits. The above simple expressions are adequate for most practical cases where L is large.

SIGNIFICANT WINDING RESISTANCE. With large L and many turns, winding resistance may sometimes become significant in relation to the load resistance, causing a drop in gain. A more accurate set of expressions is then:

$$\text{Gain} = \frac{2R_3}{k(R_1 + R_w) + \frac{1}{k} R_3}$$

$$k \text{ (opt)} = \sqrt{R_3/(R_1 + R_w)}$$

$$\text{Gain (opt)} = \sqrt{R_3/(R_1 + R_w)}$$

where R_w is the sum of primary resistance (half primary in the SPDT case) and reflected secondary resistance. In most cases, the more approximate value of k (opt) would be used, the gain being checked by the first of the above expressions.

L VERY LARGE. If L is so large that the intermediate waveform steps are close to the baseline, a further refinement can be made in the analysis. Assuming that L is infinite, but allowing for winding resistance, we see that the mean input resistance at C_1 is:

$$\frac{t_1}{2t_2}(R_w + R_3/k^2)$$

$$\text{Hence } V_1 = V \frac{\frac{t_1}{2t_2}(R_w + R_3/k^2)}{R_1 + \frac{t_1}{2t_2}(R_w + R_3/k^2)}$$

$$\text{And Gain} = 2k\frac{V_1}{V}\frac{R_3/k^2}{R_w + R_3/k^2} = \frac{2R_3}{k\left(\frac{2t_2}{t_1}R_1 + R_w\right) + \frac{1}{k}R_3}$$

$$\text{Whence } k \text{ (opt)} = \sqrt{\frac{R_3}{\frac{2t_2}{t_1}R_1 + R_w}}$$

$$\text{And Gain (opt)} = \sqrt{\frac{R_3}{\frac{2t_2}{t_1}R_1 + R_w}}$$

The gain, measured in terms of peak-to-peak output, thus increases with increasing off time, but of course the pulse width decreases. Long off

times and the consequent large values of L are generally undesirable and would only be used if there were some special requirement for this type of waveform.

OPERATION WITH SMALLER L. It is more economical to use smaller values of L. As L is reduced, the magnetizing current increases in inverse proportion, and the stored energy $\frac{1}{2}Li^2$ also increases.

Hence the circuit comprising L, R'_3 and a buffer capacitor C_b can be rendered oscillatory during the off periods. If the buffer capacitance is chosen correctly, the oscillation can be made such that the voltage exactly reaches $-V_1$ at the end of the off period. It then precisely matches the voltage on the filter capacitor when this is re-connected in the reverse

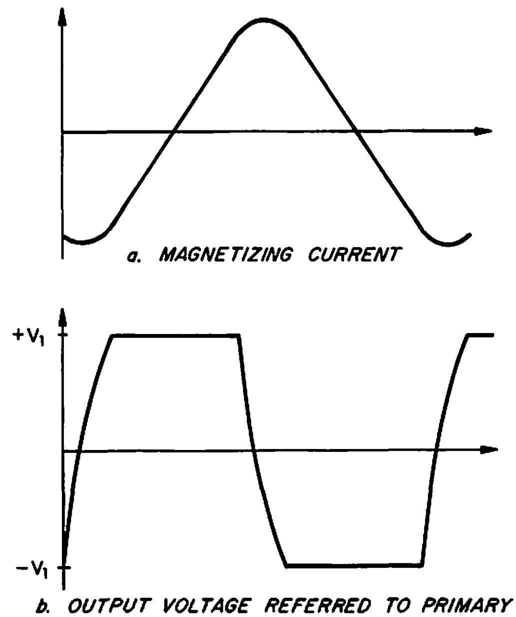


FIGURE 22. *Current and voltage waveforms — low L. Ideal buffering.*

direction. There are no losses due to transients, and the arrangement is analogous to the resonating of an inductance by parallel capacitance in sine wave circuits. Typical current and voltage waveforms are shown in Figures 22a and 22b.

If the off time is not too long, the input resistance is again approximately R'_3, and the optimum values of k and of gain are $\sqrt{R_3/R_1}$ as before. This is quite a good approximation, since an appreciable output is provided during the off time, and corresponding energy is drawn from the source. With lower values of L, winding resistances are in practice lower, and can usually be neglected in relation to the load.

The effects of under buffering (C_b too small) and of over buffering (C_b too large) are shown in Figures 23a and 23b. Both involve a loss of energy which can be shown to be proportional to the buffering error and the nominal value of C_b. In the sine wave analogy, the current drawn by an incorrectly resonated inductance would be proportional to the error and the nominal capacitance.

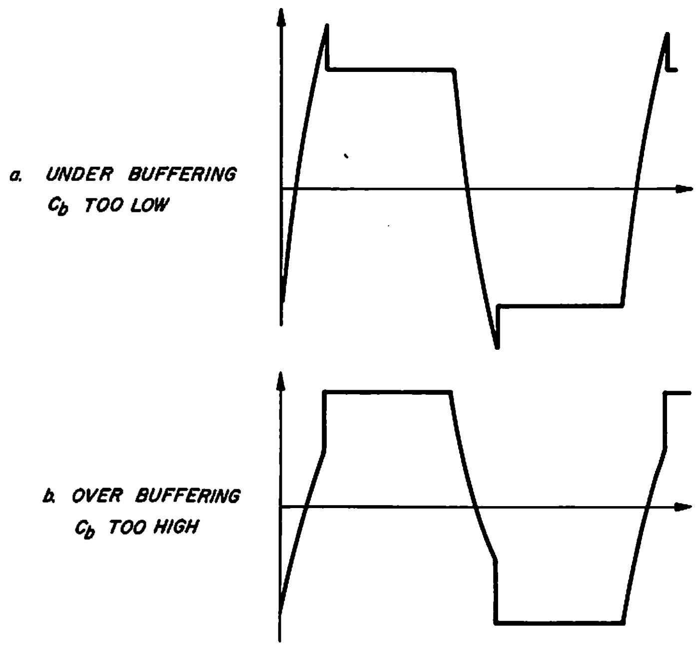


FIGURE 23. *Voltage waveforms — incorrect buffering.*

MAGNITUDE OF L AND C_b. A complete analysis of the circuit has been given by Evans[1] who shows that it is possible to choose an ideal buffer capacitor, giving just the right amount of reverse swing, provided that:

$$\frac{L}{R'_3} \leqslant \frac{t_1}{5} \text{ to } \frac{t_1}{10} \text{ approximately,}$$

for values of off time typically found. With very small off times, somewhat larger values of L are permissible. The above condition covers off times up to about 30% of the cycle. We may note that this condition takes over approximately where that for the "high L" case with no spikes left off, since $t_1/5$ is roughly equal to $t_2/2$.

The ideal value of buffer capacity is given by:

$$\frac{1}{2}\, t_2 \sqrt{\frac{1}{LC_b}} = \cot\left\{ t_3 \sqrt{\frac{1}{LC_b}} \right\}$$

[1]Evans R. H., Vibratory Power Converters. Proc. I.E.E. Monograph No. 109 R. September, 1954 (102C, p 62).

which becomes exact as L tends to zero, and is correct within about 10% provided that L/R'_3 is not greater than about $t_1/10$. This expression is depicted as a graph in Figure 24, and can be used to estimate the required value of C_b. This should always be checked experimentally, however, since transformer self capacity may be present. This is especially important if the off time is small, and L is near its upper permitted value, since both these factors tend to make the ideal C_b small.

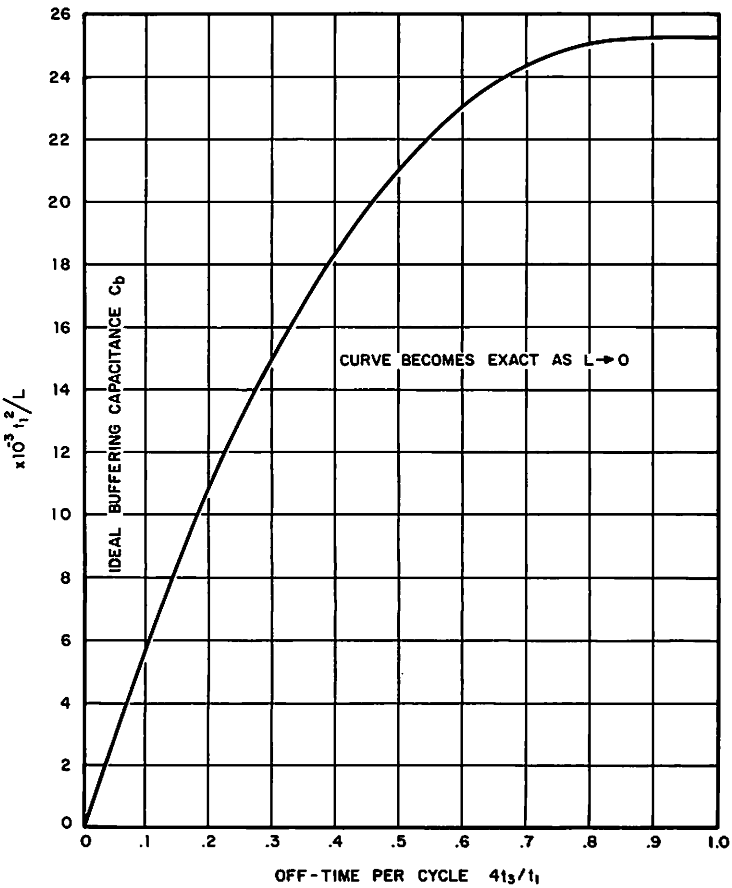


FIGURE 24. *Ideal buffering capacitance versus off time.*

Our theory so far sets no lower limit for L. But two practical factors do so, the first being winding resistance, and the second, errors in buffering.

As L is reduced, the magnetizing current increases and gives rise to greater primary losses, even if the primary resistance decreases with L. Suppose that L/R'_3 is several times less than $t_2/2$ so that we are well into the oscillatory region. Then the peak magnetizing current i_o, which equals $V_1t_2/2L$ is several times greater than V_1/R'_3, and the primary losses are

largely due to the magnetizing component of primary current. This is triangular, and the losses during one dwell period, taking $t = 0$ when $i = 0$, are:

$$2\int_0^{t_2/2} R_p\, i^2\, \delta t = 2\int_0^{t_2/2} R_p \left(\frac{i_0 t}{t_2/2}\right) \delta t$$

$$= \frac{1}{3}\, i_0^2\, R_p\, t_2 = \frac{1}{12}\, V_1^2\, R_p\, t_2^3/L^2$$

These losses become significant when they approach the energy supplied to R'_3 during the same period, namely $t_2 V_1^2/R'_3$, the gain being approximately halved when they are equal. Hence a lower limit for L is given by:

$$\frac{L}{R'_3} \cdot \frac{L}{R_p} \geqslant \frac{t_2^2}{12} = \frac{t_1^2}{48} \text{ approximately.}$$

This is still not very helpful if R_p is very low. However, the lower we make L, the larger is the ideal C_b and the greater the losses if buffering is not correct. Physically, this is because of the larger energy circulation involved. Some drift in contact adjustment is inevitable during the life of a chopper, and this upsets the buffering. Hence L/R'_3 must not be too low, and a minimum value in the region of $t_1/50$ is suggested.

If this condition is met, and R_p is made sufficiently low that $L/R_p > t_1$, then the waveform will have reasonably small droop, and primary losses will not unduly reduce the gain.

Again it is essential to test any particular design with the off time and the inductance varied through the full range of values which may be encountered, not only due to production spreads, but also due to variations over the chopper lifetime. Short off times are desirable, since they reduce C_b and the errors due to incorrect buffering.

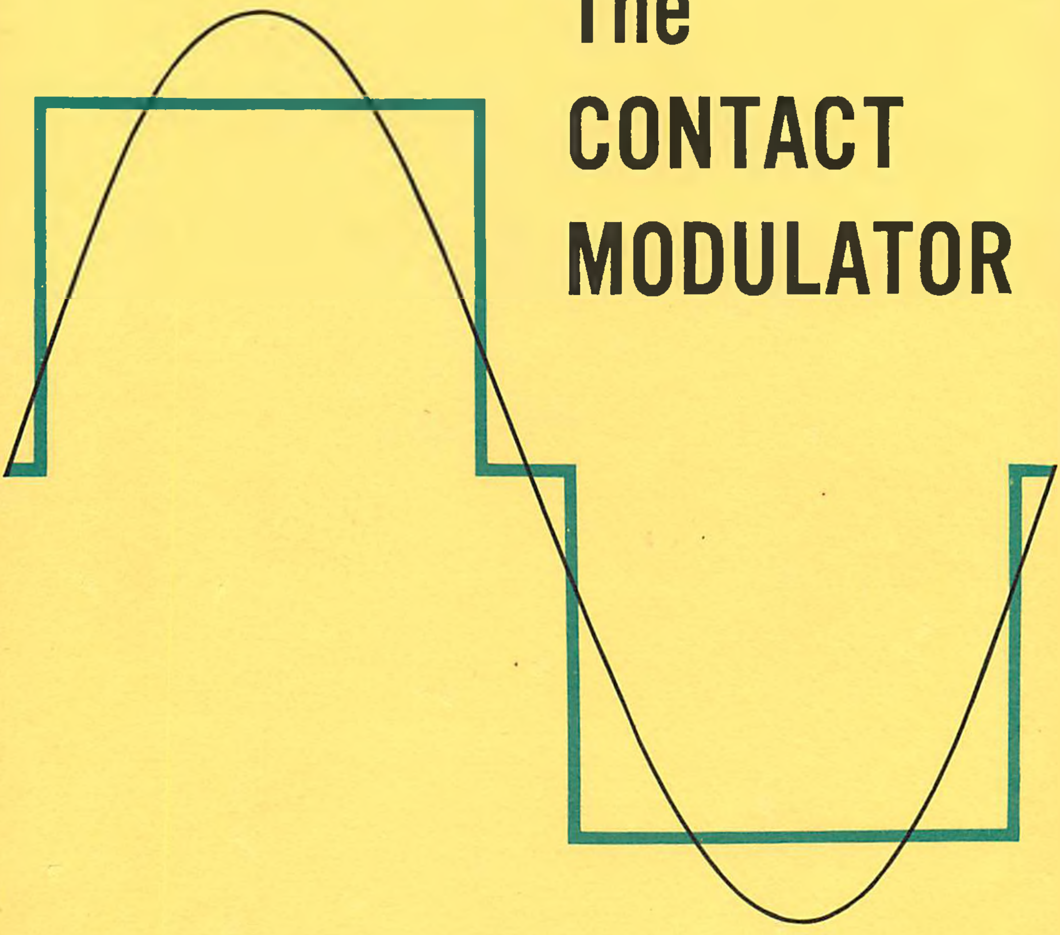
DESIGN EXAMPLE. Taking the same initial requirements as we did in the MBB example, we have:

$t_1 = 20$ m.s.
$t_2 = 9$ m.s.
$t_3 = 0.5$ m.s.
$R_1 = 1000$ ohms.
$R_3 = 100{,}000$ ohms.
k (opt) $= 10$.
Gain (opt) $= 10$.
$R'_3 = 1000$ ohms.

and the suggested range of L becomes 2 to 0.4 henrys for oscillatory conditions. This is less inductance than was required in the MBB case, and the winding resistance should be negligible in practice, so that the gain will in fact, approach 10.

AIRPAX

The CONTACT MODULATOR



PART 4: Application and Performance

THE FOURTH OF A SERIES

AIRPAX

THE
CONTACT MODULATOR

PART IV
APPLICATION
AND
PERFORMANCE

WRITTEN BY

THE ENGINEERING STAFF

OF

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TABLE OF CONTENTS

SECTION I
SOME INPUT CIRCUITS

Chopper circuits are not difficult or complex, however, experience is still helpful. There follow a number of modulator circuits, with the waveforms observed at the input of the chopper amplifier. The component values given were chosen casually for convenience rather than optimum performance. We go on to describe methods of controlling the phase relationship of contacts to chopper drive, and the manner in which chopper performance is influenced by environment.

A frequently used modulation method is the circuit of Figure 1, a half wave modulator. The input is periodically shorted to ground, and the resultant pulse is shown. The waveform shown represents negative DC input, (because the pulse is positive going), and the peak-to-peak wave height at the input grid is about the DC value at the amplifier input. Part III presents an analysis of this and other input methods.

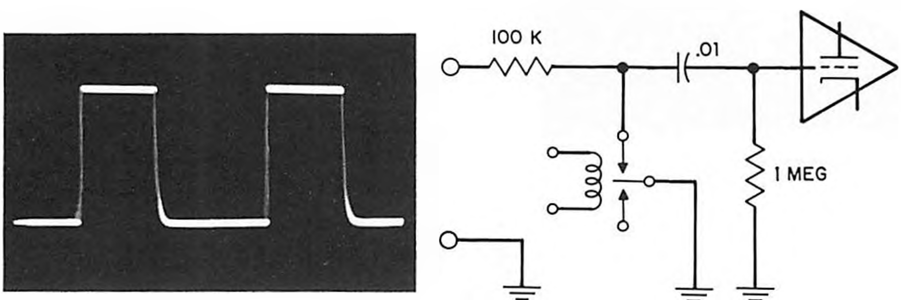


FIGURE 1. *Waveform of a half wave modulator.*

A slightly more sophisticated circuit, but one which ties up the other contact, is shown in Figure 2. (The other contact might be used for a rectifier, or to modulate another amplifier.) During the transit time of a break-before-make chopper in the circuit of Figure 2, the amplifier looks into a high impedance, the effect thereof being noticeable in the off time curvature, (discharge of C_2, stray circuit capacity). Therefore, there is some danger of noise pickup during off time, if attention is not paid to noise reduction. (See Part V, Noise in Chopper Circuits.)

A less satisfactory solution is the use of an MBB chopper, which maintains contact continuity. Of course, this makes us dependent on perfect contact action. Perfection is not always obtainable, and while make-before-break choppers are readily available, it may be better to design for maximum reliability. If the circuit is unaffected by an open contact either kind

of chopper can be used at will.

It should also be noted that with an MBB chopper, during the common time, all contacts are common and resistor *R* is grounded. If this causes undue loading on the input, it may be necessary to insert resistance in series with the grounded contact, which will in turn reduce the circuit gain. The output voltage E_o of Figure 2, will have a peak-to-peak value about twice that of Figure 1, assuming circuit components are the same in both cases.

COMPARISON. A frequent application requires the comparison of two DC values, such as a reference standard of voltage and an unknown so that the difference signal will be zero if the two values are equal. Figure 3, is a typical circuit with the amplifier input waveform shown.

The charge on the capacitor from the standard source is compared at

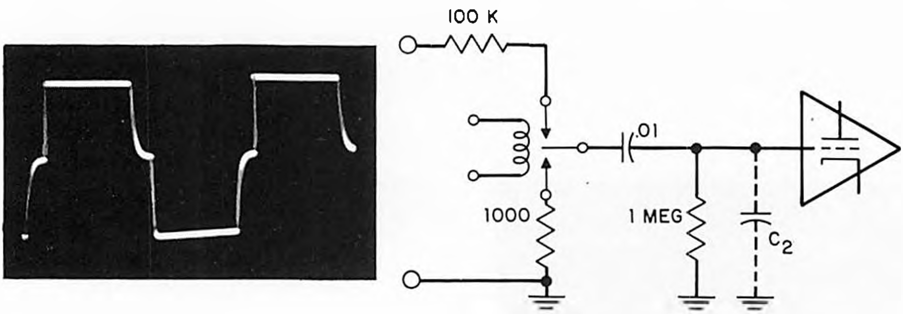


FIGURE 2. *Half wave modulator showing capacitive effect during off time.*

the grid of the input stage with the unknown DC signal to be measured, and the amplifier observes the difference between them. This is an excellent zero hunting circuit and the basis for some successful servo systems.[1, 2]

Signals need not be compared in two contacts, they can be mixed by the use of isolating resistors and examined as in Figure 4; as inputs *A* and *B* are varied the resultant patterns appear as photographed. It is important to note that Figure 4 delivers the sum of the two inputs, but Figure 3, reads the difference.

HIGH INPUT VOLTAGES. Comparison circuits sometimes are called upon to examine rather substantial voltages. Some chopper manufacturers limit certain of their models to a low voltage, probably due to contact welding or sticking if the drop across an open contact becomes excessive. Other manufacturers, AIRPAX included, permit voltages as high

[1] Digital Presentation Vacuum Tube Voltmeter. Nuut and Munsey, Electronics, January, 1956.

[2] Design of a Ratiometer. Kuehn, Electronic Equipment, November, 1955.

as 100 volts or more.

The difference [3] is one of structure. If the contacts are very small and the physical motion very minute obviously the contacts will be easily damaged. There are other problems with such structures, such as reduced resistance to extremes of temperature. On the other hand, minute motion and light contacts permit a high self-resonant frequency and operation

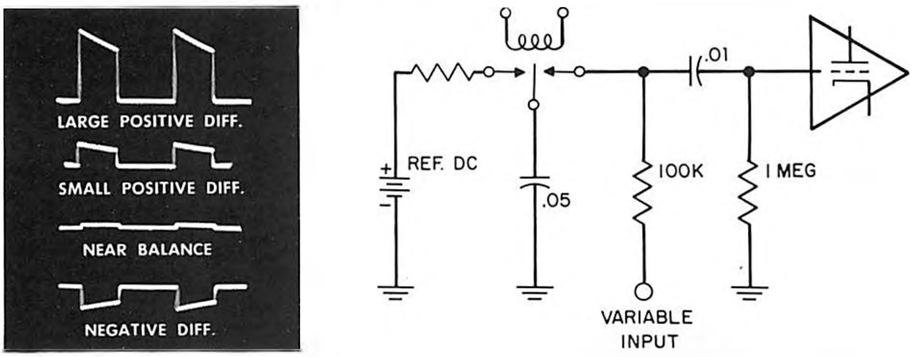


FIGURE 3. *Comparison circuit examines difference of two DC values.*

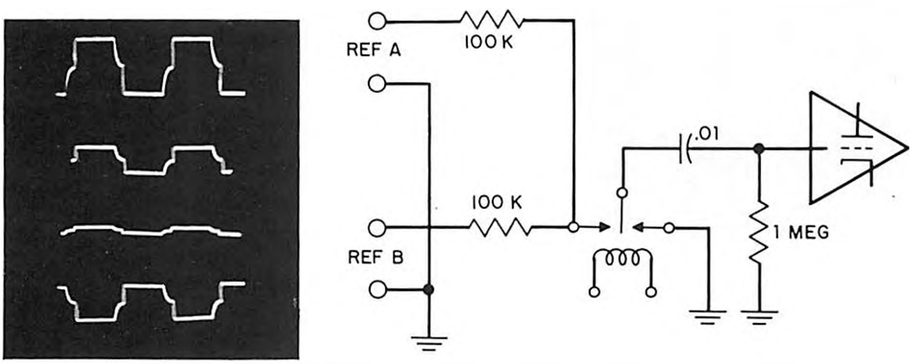


FIGURE 4. *Summing circuit adds signals A and B.*

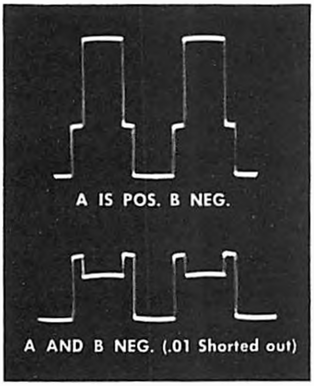
over a wide frequency range.

Still another comparison circuit is shown in Figure 5, resulting in a totally different pattern. The wave halves are substantially independent of each other, as will be observed from the waveforms. In the upper pattern one is positive, one negative; in the lower, both inputs are negative.

[3] Considerations of Relay Dynamics With An Example of Non-Linear Vibrating Reed Design, David A. Robinson, presented at Fifth National Conference on Electro-Magnetic Relays, Oklahoma Institute of Technology, Oklahoma A. & M. College, Stillwater, Oklahoma.

TIME SHARING. Figure 5, gives the possibility of using one amplifier to perform two functions. If synchronous demodulation (as for example, another chopper) is employed to restore the information obtained at inputs *A* and *B*, it is possible to perform two essentially independent operations. It should be noted that, since an AC amplifier is used, a new zero or base line is established, and without special clamping a demodulator may accept part of both signals.

INPUT ISOLATION. The transformer input circuit of Figure 6, has many virtues as was observed [4, 5] some years ago, including the ease of obtaining a floating input (both input terminals free of ground connection). It is in fact almost the only possible way, as other methods require some form of ground return. The transformer, besides providing isolation, permits considerable voltage gain. The observed waveform will contain quite a few reactive effects, as illustrated. The excessive curvature in the upper pattern is due to distributed capacity reflected into the primary. (The transformer, a Triad Type G-20 designed for choppers, had a step-up ratio of 1:8. See



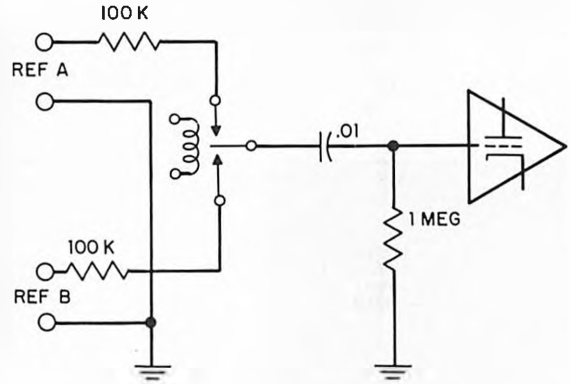


FIGURE 5. *Comparison or time sharing circuit.*

Figure 7.) In the lowest picture there appears some overshoot or "ring" at the leading edge of the wave. Reproduction of the square wave can be improved considerably by the use of AC feedback around the AC amplifier portion only; in fact, this offers the opportunity of achieving a high order of linearity without DC feedback and consequent loss of the input "float."

This circuit also permits, by careful attention to input balance, a high order of rejection to a signal between both input terminals and ground, reputedly reaching as high as a million to one. This is the "common mode rejection" given frequently as a virtue for data logging amplifiers, and is

[4] A Contact Modulated Amplifier to Replace Sensitive Suspension Galvanometers. M. Liston, et al, Review of Scientific Instruments, Vol. 17, No. 5, May, 1946.

[5] D-C Amplifier Stabilized for Zero and Gain, Williams, Jr., Tarpley and Clark, AIEE Transactions, 1948, Vol. 67.

discussed in later Parts. A typical application is a long line run from a thermocouple delivering only a few millivolts. Strong AC fields may produce signals of many times that figure, but from both lines to ground. If impedances to ground are balanced the noise signal cancels.

The transformer design must be carefully watched if a very wide range of input levels will be encountered. The waveforms of Figure 8, were taken with the circuit of Figure 6, but with DC input levels of about 200 millivolts, upper picture, and about 100 microvolts, lower picture. The slope (droop) of the wave is caused by insufficient primary inductance.

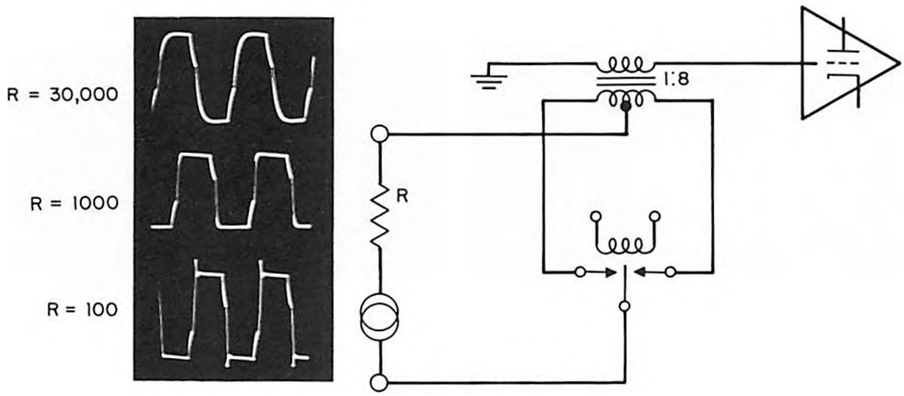


FIGURE 6. *Balanced input using a transformer; waveforms are for 3 values of source resistance.*

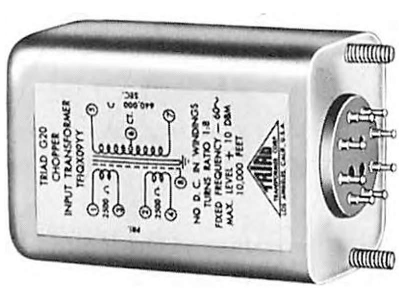


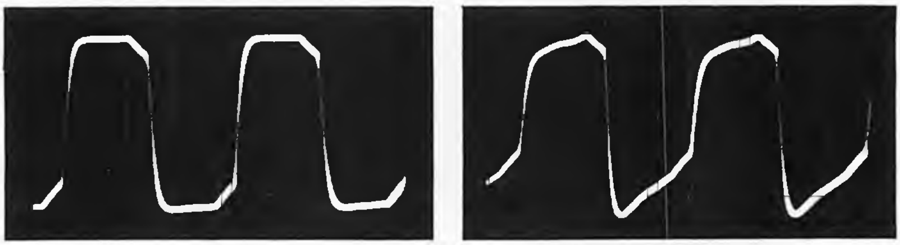
FIGURE 7. *Transformer designed for chopper use.*

Since only the operating level was changed, this means there was a reduction of permeability in the core of the transformer. High nickel laminations are needed to provide good "initial" permeability.

HIGH IMPEDANCES. A good quality capacitor is required at the input grid of most chopper amplifiers. Any leakage will be chopped and

result in noise, i.e., unwanted signal. There are times, however, when very high impedance circuits are desired and the signal levels are also rather high. Electrometer tubes, such as type VX-55, have exceptionally low grid current and can be direct-coupled in the manner of Figure 9, for special applications, for example, as a detector for an insulation resistance bridge. An application will be described in a later chapter.

TRANSISTOR AMPLIFIERS. Transistor input circuits [6] for chopper amplifiers are not different, being current operated at the input, as contrasted to voltage operation at a vacuum tube grid, they require lower impedances. So far as the presently illustrated modulation circuits are



200 μV INPUT   100 μV INPUT

FIGURE 8. *Loss of inductance in transformer at low input levels.*

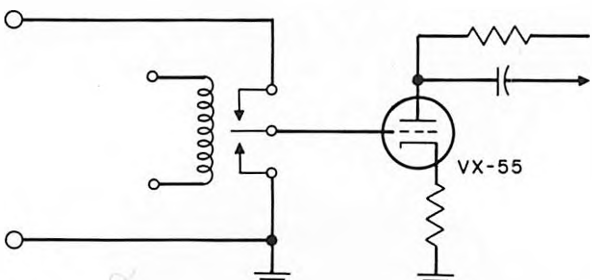


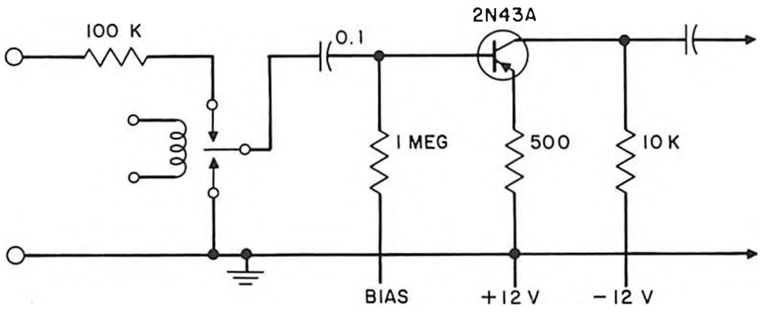
FIGURE 9. *Direct coupling has some value at very low grid current in high impedance circuits.*

concerned, there is little difference, and an input base can be substituted for an input grid in almost every instance, except that resistance values should often reduce and capacitor values increase. Later chapters present transistor circuitry in greater detail. Figure 10, illustrates two transistor input circuits.

TRANSISTOR CHOPPERS. The switching transistor can be used as a modulator, if null stability in the millivolt range can be tolerated. The transistor chopper by itself, Figure 11, is quite small, but maximum per-

[6] A Transistor D-C Chopper Amplifier. Burton, Electronic Engineering, August, 1957.

formance requires a square wave drive. AIRPAX type 6010 is illustrated, and a modulator circuit employing this chopper is shown in Figure 12. The waveforms observed with this circuit appear in Figure 13. A 400 cycle sine wave drive was used.



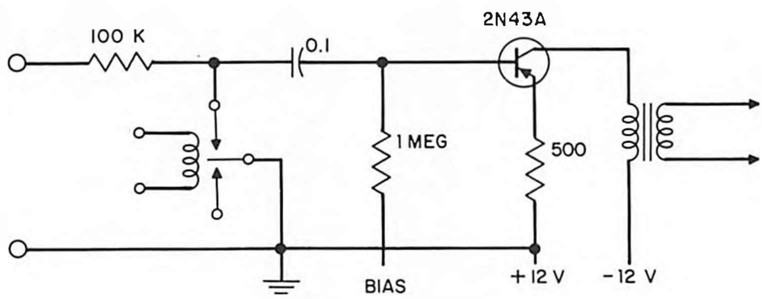


FIGURE 10. *Modulation circuits for transistors are substantially the same.*

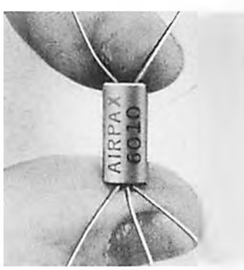


FIGURE 11. *Transistor choppers are intriguingly small and light.*

Probably the most serious fault in the performance of a transistor chopper is its susceptibility to change with temperature, which is likely to have considerable effect on both null stability and gain. However, all things are relative. While a mechanical chopper has better gain, null stability and linearity by several orders of magnitude, and while the reliability of a mechanical chopper is better than the present state of the transistor

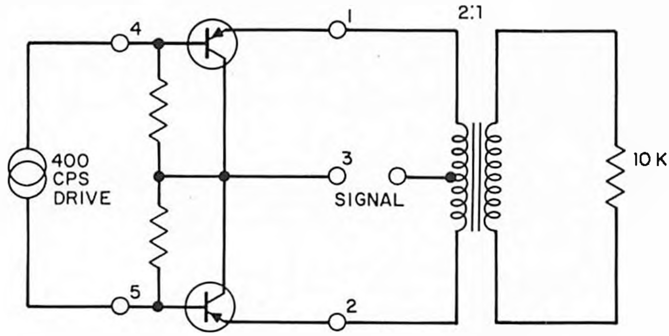
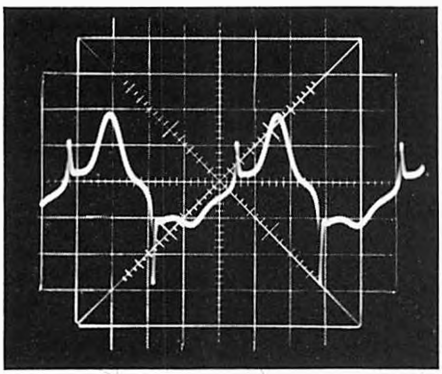
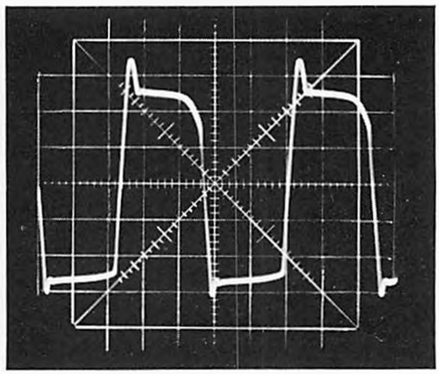


FIGURE 12. *Modulator circuit employing type 6010 transistor chopper.*



1 mv/div. vert., 0.5 msec/div. hor.
(a) Output at null



0.5 v/div. vert., 0.5 msec/div. hor.
(b) Output with signal

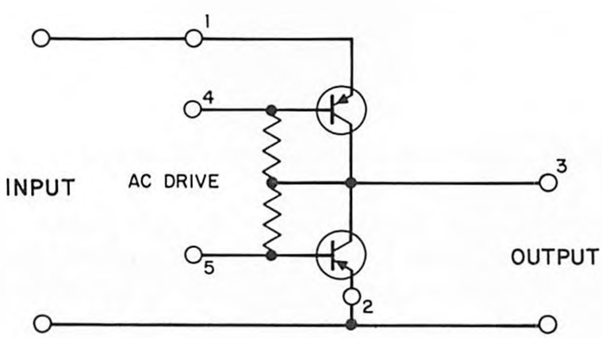


FIGURE 14. *Series-shunt arrangement of transistor chopper.*

art permits, still the transistor chopper is very small and light, and if null, gain stability and linearity are not critical the reliability is no worse than other transistor devices. The signal voltage must of course be limited to avoid transistor damage.

Another modulator arrangement appears in Figure 14, and a modulator-demodulator circuit is shown in Figure 15, both of which are quite similar to previously described devices using mechanical choppers. The transistor chopper should be fully saturated and it is usually necessary that

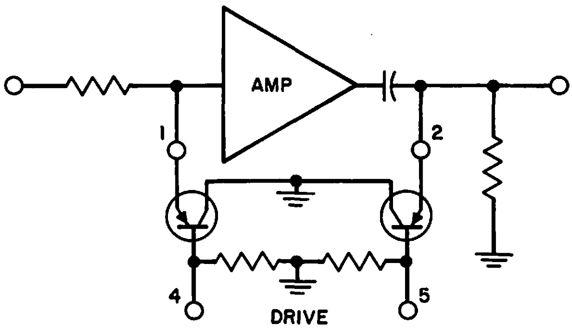


FIGURE 15. *A modulator-demodulator arrangement.*

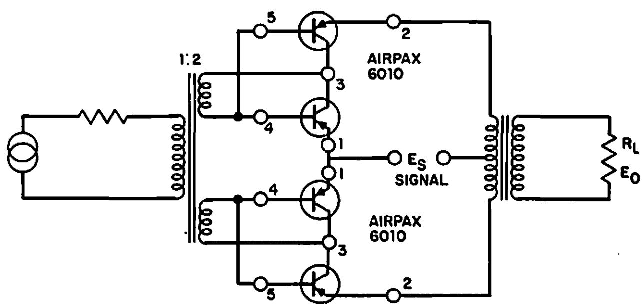


FIGURE 16. *Full wave modulator improves the null stability.*

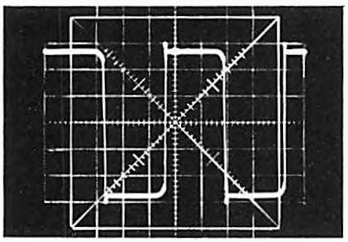
the drive voltage considerably exceed the maximum signal voltage. A square wave drive makes possible lower null voltages and also permits better performance at high signal levels. Many other[7] circuit arrangements are possible.

One of the better circuit arrangements appears in Figure 16, using two type 6010 choppers. This provides a much better null, or minimum offset, typical values being in the vicinity of 100 to 200 microvolts at room temperature and about 500 microvolts at 85°C. The linearity is about 1%, down to about 10 millivolts, and begins to get much worse as signal and

[7] Application Notes For Transistor Choppers, available from AIRPAX ELECTRONICS, INCORPORATED, Cambridge Division, Cambridge, Maryland.

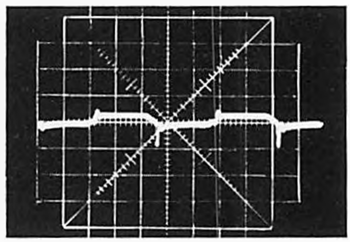
offset values become equal. Linearity, as well as offset, will be affected by temperature. Figure 17, shows a typical waveform resulting from the circuit of Figure 16; Figure 18, is the waveform of the null or no signal condition using the same circuit.

It will be noted that the driving transformer requires two secondaries. It is possible to package the transformer and the matched pair of transistors in a size about equivalent to an AIRPAX type 300 chopper, and AIRPAX furnishes a plug-in transistor chopper substitute. The transistor chopper is of course, not as versatile, and it is recommended that the factory be contacted with the required application.



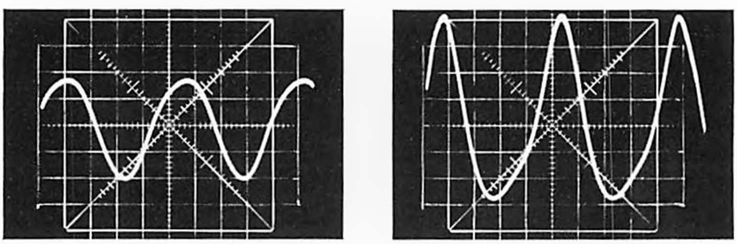
$R_L = 10$ K $E_S = 1.5$ v
0.5 v/div. vert., 0.5 msec/div. hor.

FIGURE 17. *Signal voltage at E_0 has this waveform.*



$R_L = 10$ K $E_S = 0$
1 mv/div. vert., 0.5 msec/div. hor.

FIGURE 18. *Oscillogram of null value of 0.19 millivolts.*



$R_L = 10$ K $E_S = 100$ mv
50 mv/div. vert., 2 μsec/div. hor.

$R_L = 10$ K $E_S = 0$
10 mv/div. vert., 2 μsec/div. hor

FIGURE 19. *With 100 KC drive in circuit of Figure 16; (a) E_o output of 46 millivolts, (b) null value of 15.7 millivolts.*

The phase relationship of the transistor chopper is quite close to the driving voltage at the lower frequencies. Operate and recovery time, plus the effects of capacity, result in phase lags becoming noticeable in the vicinity of 10 kc. The AIRPAX type 6010 can be driven as high as 100 kc, but at that frequency performance has deteriorated considerably, as is evident from Figure 19.

OTHER MODULATORS.[5] A complete description of the many modulation methods is beyond the scope of this text. Almost innumerable varia-

[5] D-C to A-C Modulators, George Sideris, Electronics, p. 47, January 23, 1959.

tions occur to the fertile minds of engineers involved in DC systems.

Magnetic modulators offer some promise, being quite stable and possessing a magnificent reputation for high reliability. However, if one is sufficiently concerned with reliability to need the probable 100,000 hours [9] of life obtainable from magnetic devices, it seems sensible to use a magnetic amplifier rather than transistor or tube types with magnetic modulation.

The mechanical chopper out-performs tubes and probably transistors, with regard to life and general reliability. Magnetic modulators of fundamental output are necessarily of rather low impedance and have null stabilities not usually as good as the best transistor chopping circuits. Figure 20, shows the circuit of a "fundamental" type of magnetic modulator. Another limitation is slow response time, which gets slower as sensitivity improves.

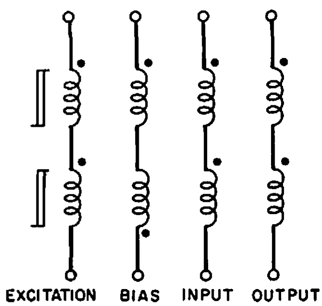


FIGURE 20. *"Fundamental" type magnetic modulator. Response is slower as sensitivity is increased.*

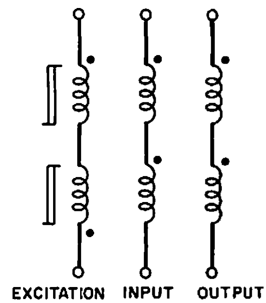


FIGURE 21. *"Second Harmonic" type magnetic modulator. The frequency response is limited.*

Second harmonic magnetic modulators exhibit better performance, probably better than most transistor choppers. The limitations of low impedances and limited frequency response are still present. A typical circuit is shown in Figure 21. Now, we can beat the response time rap by the stabilized amplifier techniques described by Goldberg [10] and others, as discussed in later chapters. However, it is somewhat illogical to refuse a transistor chopper in a transistor amplifier, and as pretty often repeated in this text, the mechanical chopper may not be too reliable, but 5,000 hours of life is considered normal and much longer periods have been measured.

The null stability of magnetic amplifiers can be pretty good, even at moderately high impedances.[11] Figure 22, shows an AIRPAX data ampli-

[9] Bulletin 221, describing FERRAC Magnetic Computing Amplifiers, available from AIRPAX ELECTRONICS, INCORPORATED, Seminole Division, Fort Lauderdale, Florida.

[10] Stabilization of Wide-Band Direct-Current Amplifiers for Zero and Gain, Goldberg RCA Review, p. 296, June, 1950.

[11] A Low-Level, High-Accuracy, D-C Magnetic Amplifier, Blas A. Mazzeo, Electrical Manufacturing, November, 1958.

fier. Measurements of a typical amplifier show nulls within +10 and −20 microvolts over 100 hours and −55°C to +85°C. The specification limit is ±100 microvolts. The input impedance is 50,000 ohms at a gain of 100, reducing to 10,000 at a gain of 500.

CAPACITIVE CHOPPER. A useful device for the investigation of small electrostatic charges may be built as shown in Figure 23. This device is essentially a capacitive "chopper" input circuit followed by a "floating" cathode follower.

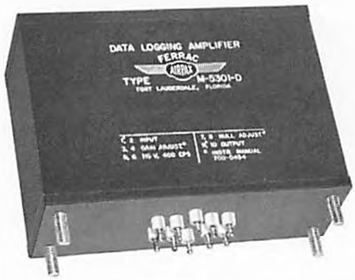


FIGURE 22. *Magnetic Data Logging Amplifier has input impedance of 50,000 ohms at a gain of 100; variable gain from 100 to 10,000; complete input isolation.*

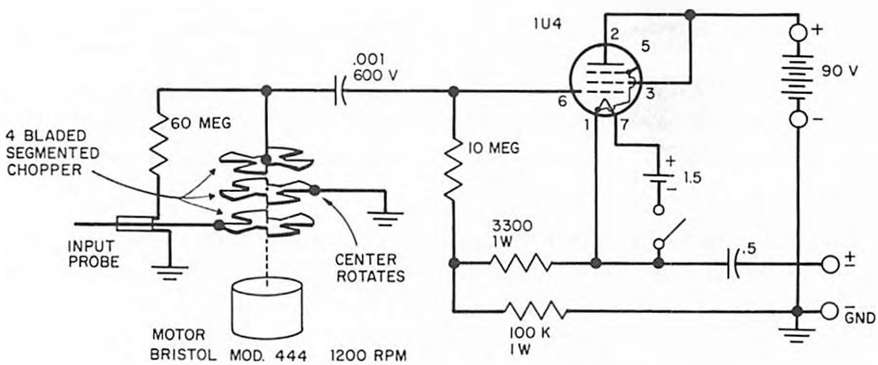


FIGURE 23. *Circuit diagram of electrometer modulator.*

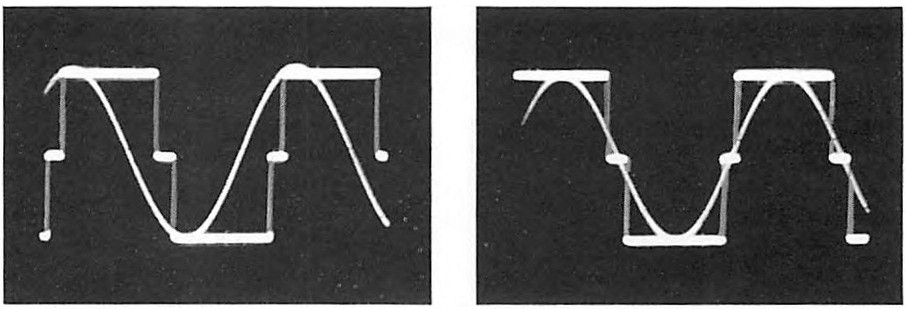
The output, which consists of essentially rectangular pulses at 80 cycles/second, gives an indication of input polarity by changing phase. These pulses may be viewed on an oscilloscope. The sensitivity of this device is such that the determination as to the anionic or cathionic nature of a plastic may be determined by rubbing the material against the probe.

Needless to say, the high input resistance of the device necessitates the use of Teflon or Polystyrene insulation in the input circuit. Such a commercial instrument may be purchased from the Pacific Transducer Corp, 11836 West Pico Blvd., Los Angeles 64, California.

SECTION II

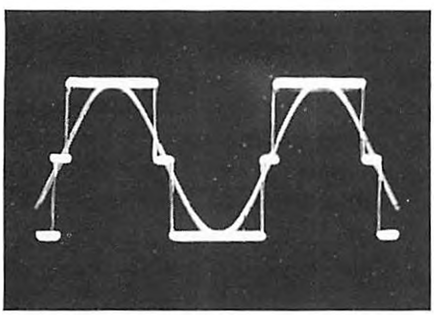
PHASE RELATIONSHIP

Having obtained modified AC into an amplifier, it becomes necessary to use it. Often, as for operational amplifiers, the signal is rectified after amplification, then filtered to restore DC. Sometimes the signal controls a motor, either directly, or by timing the firing of thyratrons. This, and also the use of choppers as rectifiers, makes the phase angle of interest. The discussion following is aimed at mechanical choppers. Phase adjustment may of course become necessary for solid state devices to accommodate transducers or output devices.

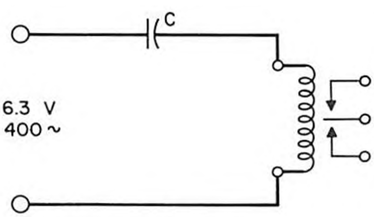


65° LAG | 0°

FIGURE 24. *Phase angle relationship to drive can be adjusted readily.*



0°



CHOPPER TYPE	C (IN MFD)
300	1.33
360	1.33
370	1.6

FIGURE 25. *Introduction of series capacity permits operation at zero phase angle.*

In any mechanical chopper there exists two components of phase lag, the L/R relationship of the drive coil, and an electro-mechanical lag of the armature. Where chopper phase angles are of interest, 0° and

90° are most popular for obvious reasons, and 0° is the most easily obtained by external capacity. Introduction of series capacity introduces a leading current, and it is quite practical to correct most choppers to zero. Figure 24, illustrates a type 300 chopper photographed in relationship to its driving signal, (a type 175 chopper operating at 60 cycles provided the time sharing arrangement to permit the superimposed scope pictures).

At 6.3 volts a simple series capacitor is used to adjust phase, as in Figure 25. The capacitor values are given for several 400 cycle choppers, (this is not practical at 60 cycles). At higher voltages a series-parallel RC combination becomes preferable, as in Figure 26.

There is an infinity of possible combinations, and optimum values for various drive voltages, such as 115 volts, are given in Figure 27. Now, a zero phase angle is not necessarily the best value, due to other phase shifts which probably exist in amplifiers, etc., but these values will provide a good beginning. They are recommended by the factory as providing best

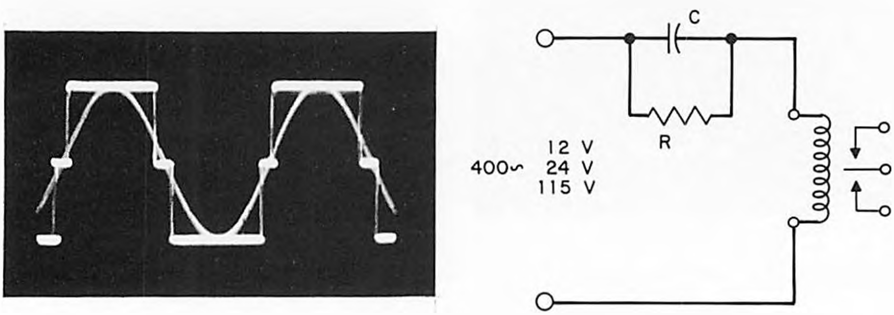


FIGURE 26. *Series-parallel combination provides optimum performance.*

performance. One or two volts are required to "start" a 6 volt chopper. Most AIRPAX choppers have substantial motion, large contact area and high contact pressure; indeed, that is one of the reasons for the fact that AIRPAX has greater reliability than any other chopper supplier. There is a drive impedance change, between start and run; like any other motor, the starting current is higher. Therefore, when designing phase changing networks, see that adequate starting current is supplied; if in doubt, ask the factory for a recommendation. It should be remembered that a check on only one or two choppers may not yield optimum results if production quantities are being considered. For network operation from 115 volts, or other values higher than 6.3 volts, the circuit constants are usually adjusted to provide from 7½ to 9 volts across the chopper coil.

Over wide temperature extremes the copper resistance of a chopper changes considerably. The phase stability is improved by a constant current drive, like that of Figure 28. The phase angle of the chopper will be

CHOPPER TYPE	TOTAL PHASE ANGLE (DEGREES)	ELECT. PHASE (DEGREES)	MECH. PHASE (DEGREES)	DRIVE CURRENT MA	COIL Z (OHMS)	115 VOLT NETWORK		24 VOLT NETWORK		12 VOLT NETWORK	
						R OHMS	C mfd	R OHMS	C mfd	R OHMS	C mfd
300	65	30	35	22.5	280	4300	.061	1200	.33	1100	.68
350	65	30	35	24	262	6800	.050	1200	.33	1200	.68
360	55	30	25	22.5	247	4300	.050	1200	.33	1000	.68
370	55	24	31	26	242	5100	.050	1000	.39	840	.94
600	65	32	33	50	126	2600	.10	560	.68	430	1.68

FIGURE 27. *Circuit of Figure 26 permits zero phase angle when RC network uses values listed, at 400 cycles.*

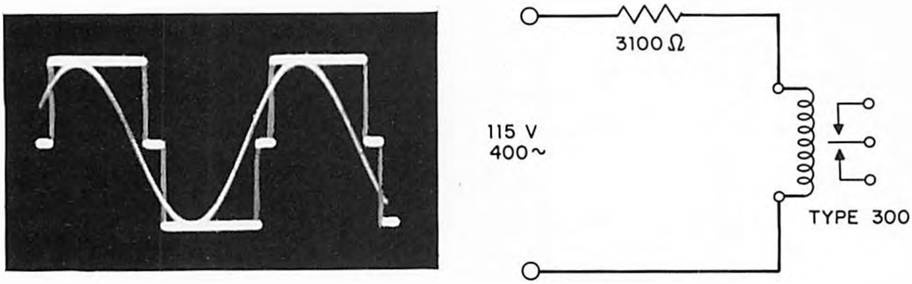
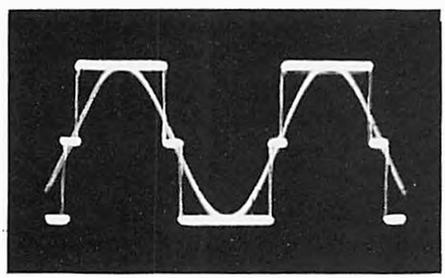
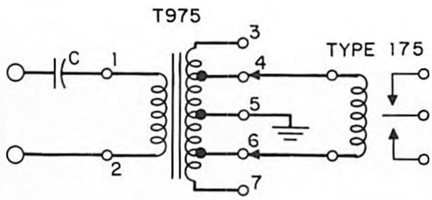


FIGURE 28. *Constant current source produces angle of about 35°.*





	C (IN mfd)	CONNECTIONS
ONE CHOPPER	.250	4 - 6
TWO CHOPPERS	.540	3 - 7

FIGURE 29. *Transformer permits zero angle and grounded center-tap on drive.*

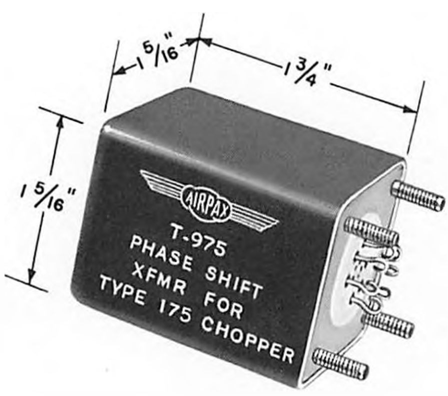


FIGURE 30. *T975 is a small transformer.*

close to the "mechanical" angle, since compared to the inductive reactance at 400 cycles, the circuit resistance becomes high, the current lag becoming negligibly low.

It may be desirable to shift a phase angle and still permit a balanced drive (to keep noise at a minimum). Figures 29 and 30, describe an AIRPAX type T975 transformer for use with type 175 choppers at 60 cycles. The exciting current of the transformer primary affects the angle, as also does the chopper load, and the primary inductance of the transformer becomes important. Type T975 transformers, tested with type 175 choppers for optimum performance at zero angle, are stocked by AIRPAX at the Cambridge Division.

SECTION III
THE AMBIENT CHANGES

Engineers become concerned with component changes under extremes of temperature, altitude, drive frequency, vibration, humidity, shock, dust, fungi growth, and other deleterious assaults of man and nature. Being hermetically sealed in metal and glass, an AIRPAX chopper is resistant to altitude, dust, fungus and humidity.

Humidity could, during prolonged exposure and while carrying substantial DC polarization, cause electrolytic tracking across glass. This is an old story in transformer feed-through leads; fortunately choppers seldom carry sufficient DC potential to reduce the insulation resistance of glass enclosed terminals appreciably.

The two parameters of a chopper most likely to be affected are dwell time and phase angle. Noise is not usually changed, again due to hermetic sealing. It should be noted that contamination and moisture on the socket

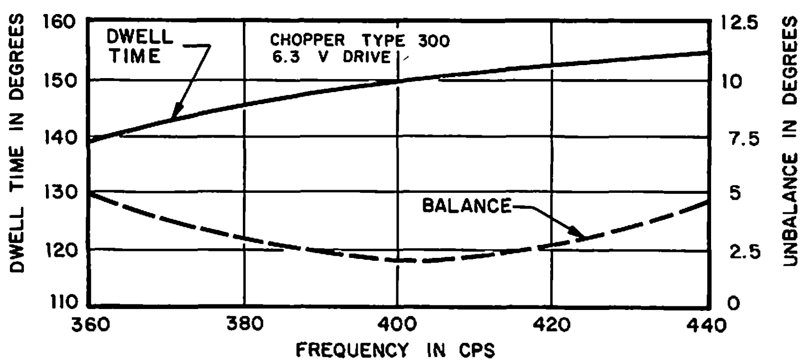


FIGURE 31. *Variation of dwell time and balance with driving frequency.*

or external connections are quite likely to cause noise, substantial values, in fact, from the presence of electro-chemical potentials. See Part V, "Noise in Chopper Circuits."

Dwell time is frequently important, as a change in dwell time will be effectively a change in gain in some circuits. Phase angle is not important in many DC amplifiers, except for the relation between modulator and demodulator. (Two AIRPAX choppers of the same part number will remain almost identical with each other, even while they change from a change of environment. Several users have taken advantage of this fact to provide as high as 4 pole double-throw operation, by simply operating units in parallel.)

ENVIRONMENT AND THE DWELL TIME. Figure 31, illustrates the change in dwell time with driving frequency of a production model

type 300 chopper. Specification limits are usually 380 to 420 cycles. The dwell time increases with frequency, but the chopper remains balanced, which is to be expected if the dwell times are initially symmetrical. Figure 32, is another typical curve, the effect of varying the coil voltage from 3 to 9 volts (5.9 to 6.7 is a more usual range to be expected). Figure 33, illustrates the relative indifference of the chopper to temperature, over the extreme of —65°C to +125°C. Again specification limits are less than the capability of the chopper, which normally operates from —50°C to +85°C.

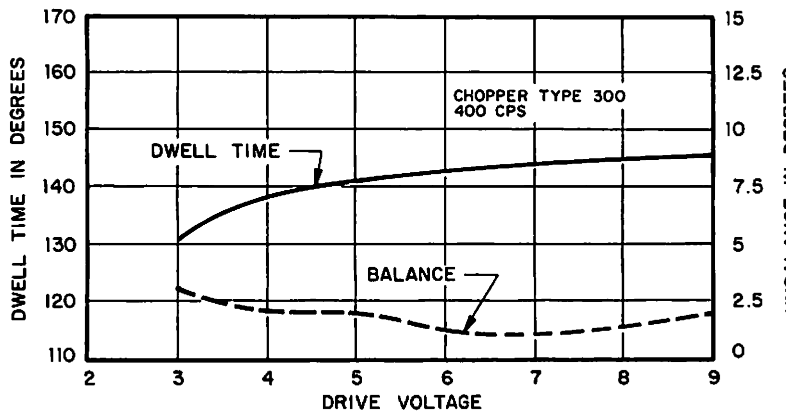


FIGURE 32. *Variation of dwell time and balance with driving voltage.*

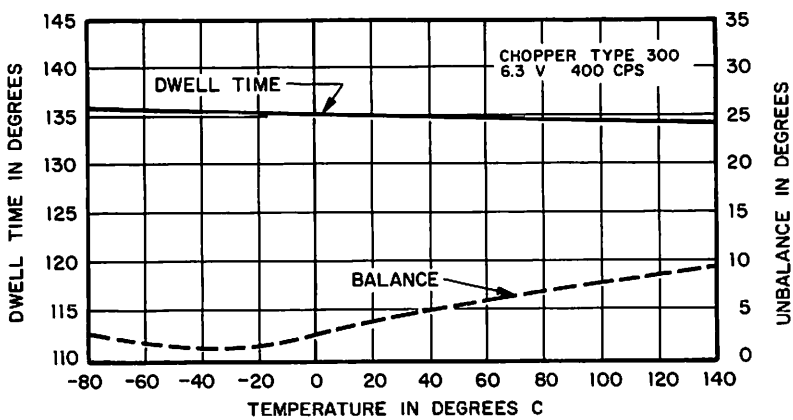


FIGURE 33. *Temperature effect on type 300 chopper.*

Figures 31, 32, and 33, are values averaged from large groups, accordingly are illustrative of what can be expected at 400 cycles, at least from an AIRPAX type 300 chopper. Figure 34, is the measured dwell time during 25,000 hours of life testing (2.86 years of operation). The probable spread of dwell time figures is well illustrated in Figure 35,

showing the extreme range of dwell time as measured on 20 type 300 choppers during 1,500 hours of life testing. Complete measurements were made every 250 hours and the dwell time was recorded automatically every few hours. (Refer to Part II, Measurement.)

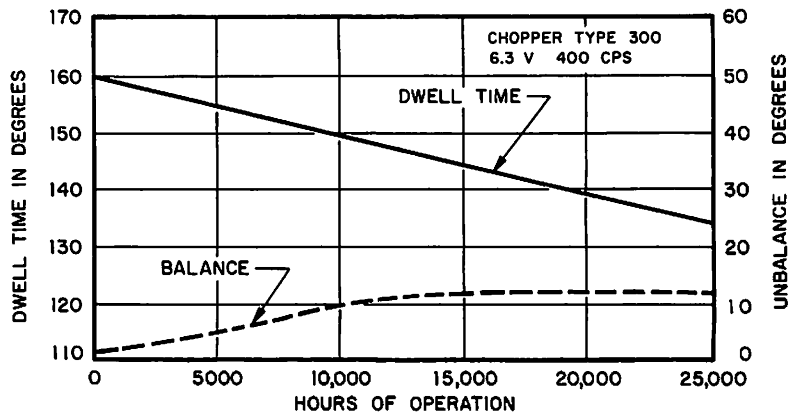


FIGURE 34. *LIFE TEST DATA. Chopper operating continuously in "dry" circuit. Dwell time recorded every 6 hours; a complete test performed every 100 hours.*

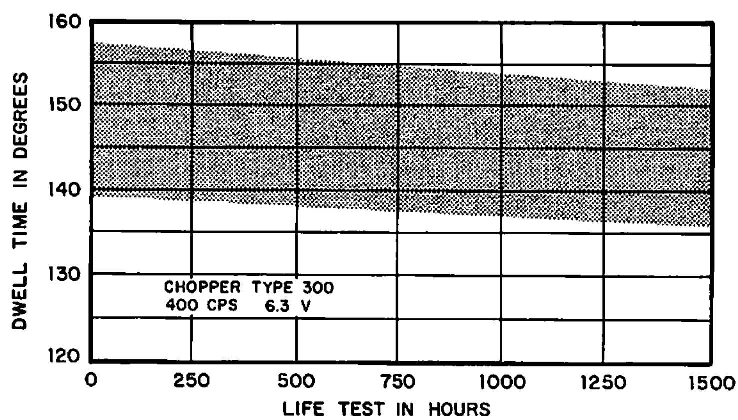


FIGURE 35. *Maximum and minimum dwell times recorded on 20 type 300 choppers selected at random from production.*

THE PHASE ANGLE VARIATION WITH AMBIENT. The phase lag of a signal from chopper contacts is composed of two elements, the L/R relationship of the coil plus a "mechanical" lag of the armature. As the operating frequency increases the phase angle also increases, and the relationship is plotted in Figure 36. An increasing voltage reduces the

angle, as illustrated in Figure 37. When operated from a ship-borne alternator, the increase in voltage and frequency may occur together, if so there will be a tendency to neutralize each other. While the plot is given over 360 to 440 cycles, and 3 to 9 volts, a chopper is not often required to operate over such extremes, as discussed earlier. When operated over a wide range of ambient temperature a phase angle change, Figure 38, occurs which is, to a large extent, due to the change in the coil resistance.

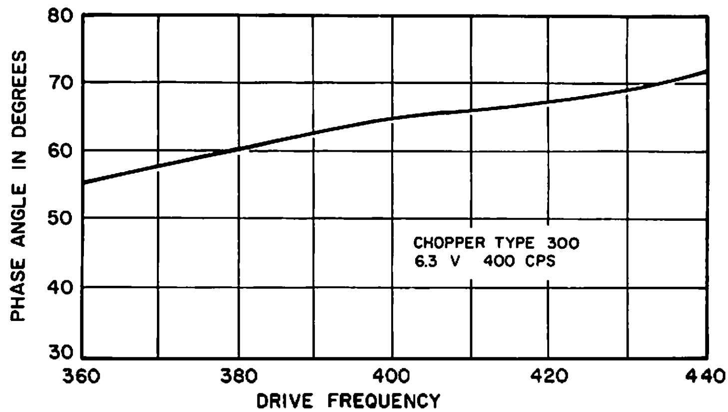


FIGURE 36. *Phase angle variation with driving frequency.*

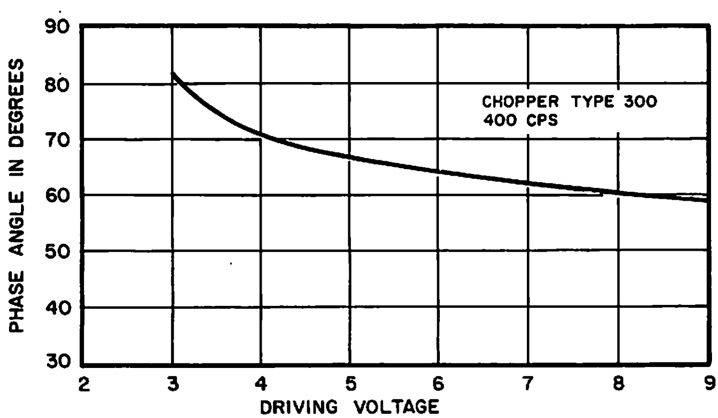


FIGURE 37. *Effect of driving voltage change.*

Between —65°C and +125°C the coefficient of copper will vary the resistance of a type 300 chopper from 110 ohms to 236 ohms. Since the reactance of a type 300 is about 150 ohms at 400 cycles, the effect of temperature on the coil time constant will be appreciable.

Figures 36, 37, and 38, are values averaged over many choppers, and two AIRPAX choppers of the same part number will "track" very closely

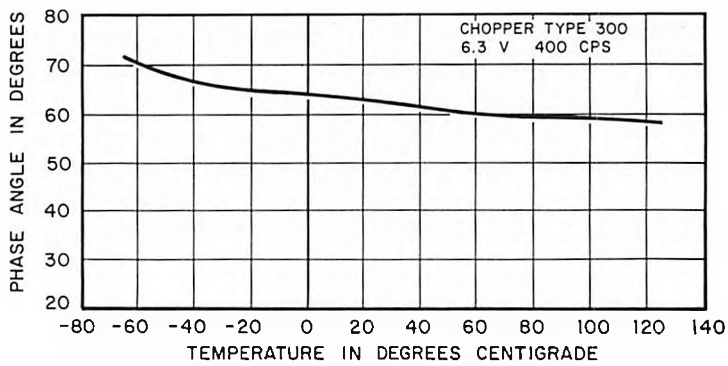


FIGURE 38. *Phase angle change with temperature is largely due to change in coil resistance.*

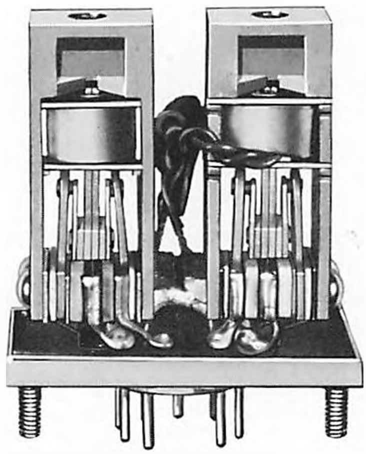


FIGURE 39. *Two choppers track together so well that they provide an excellent DPDT chopper.*

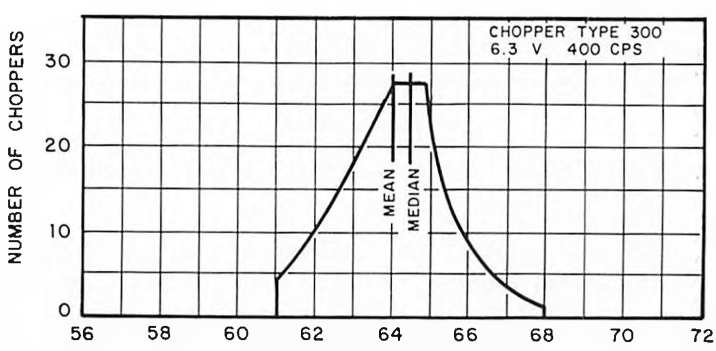


FIGURE 40. *Distribution of phase angle, 100 production choppers.*

as ambient conditions are varied. Advantage is taken of this fact to provide double-pole double-throw operation. It will be immediately obvious only to a chopper engineer, but it is probably easier to obtain accurate multiple contact operation with electrical linkage than with a mechanical coupling. Figure 39, shows a DPDT chopper employing this principle. Uniformity of production choppers is illustrated by Figure 40, a plot of the distribution observed on 100 production choppers.

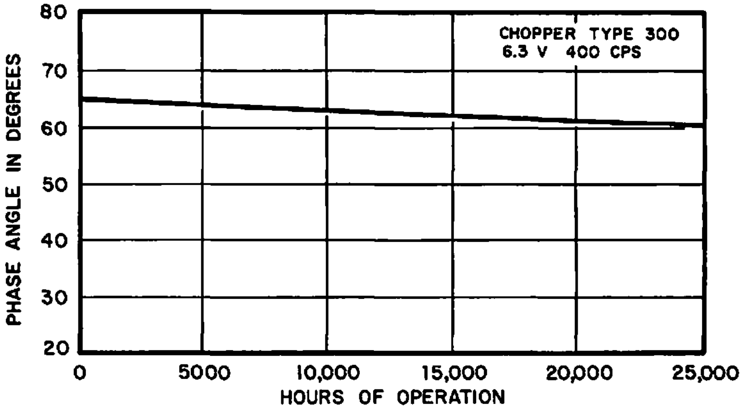


FIGURE 41. *When operated in a dry circuit a type 300 chopper maintains a nearly constant phase angle.*

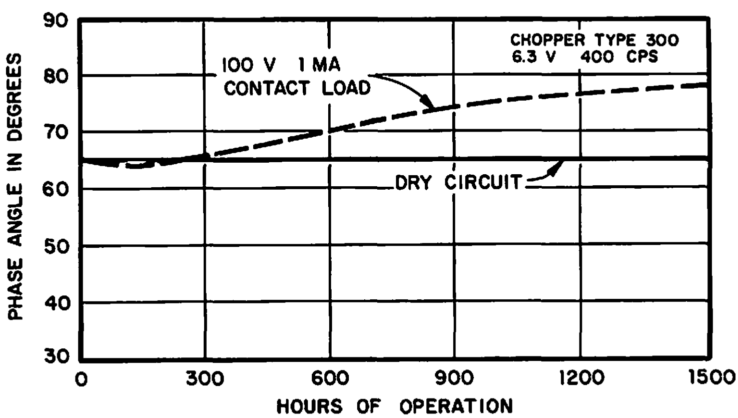


FIGURE 42. *The effect of handling moderate power on a standard chopper is to cause contact wear.*

When the phase angle becomes important, the effect of wear and changes during life become important. Figure 41, shows the almost negligible changes during practically three years of continuous "dry circuit" operation of a type 300 chopper at 400 cycles. The dissipation of moderate amounts of power, such as interruption of 1 milliampere DC at 100 volts with resistive load, will cause a gradual contact wear. The effect of this is

to increase the phase angle due to the increasing motion required, as plotted in Figure 42.

Some of the phase adjusting networks will provide phase stability over a very wide range of operating frequency, as plotted in Figure 43.

VIBRATION PROBLEMS. Chopper type 300 and many others, employ an armature hinged at one end. If the acceleration forces applied to a chopper are sufficiently high there will result what is termed contact "derangement", essentially a modulation of the phase angle.

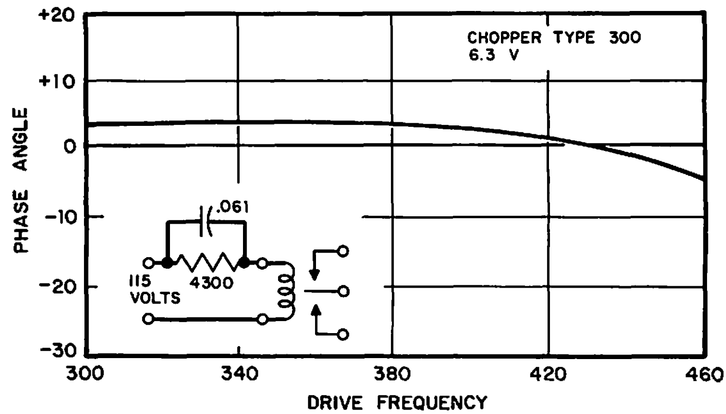


FIGURE 43. *The use of networks improves the phase stability.*

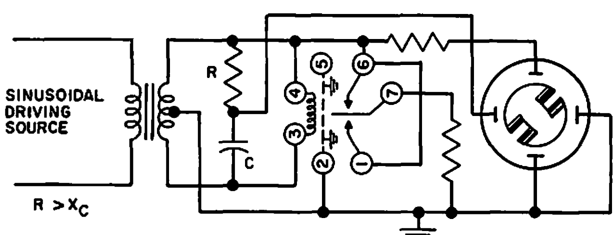


FIGURE 44. *Contact derangement as it appears on a polar scope presentation.*

The effect of vibration on chopper action is measured in electrical degrees of contact derangement. Contact derangement is the aggregate fluctuation in make or break due to bounce, chatter, phase modulation, and dissymmetry due to vibration (see Part II). Figure 44, shows these characteristics on polar coordinates as they appear in a simple test circuit.

Type 300 will survive vibration testing without damage, but as noted in Figure 45, at frequencies approaching 400 cycles, the acceleration forces exceed the armature drive forces. Mechanical self-resonance occurs

at about 550 cycles, at vibration frequencies in this vicinity the acceleration forces take over and the armature ceases to follow the drive signal.

In Series 370 choppers, the armature is supported on a pivot. As a result, external mechanical vibration imparts only translation to the armature — the same as it does to the fixed contacts. The balanced structure effectively isolates the armature from extraneous torques that could interfere with its desired motion or displace it relative to the fixed contacts.

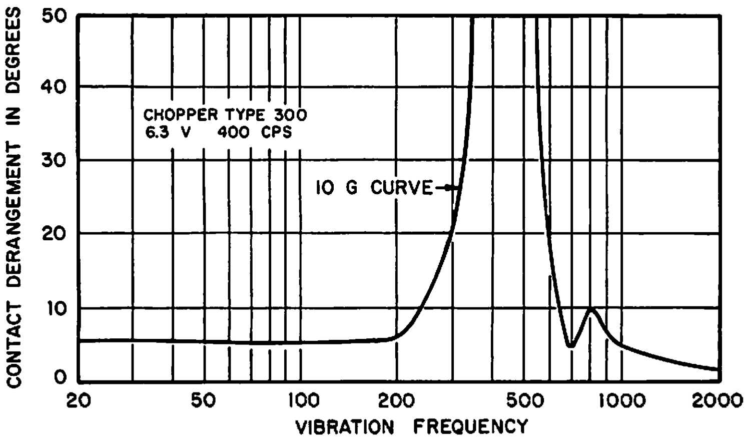


FIGURE 45. *Type 300 choppers are not damaged by high frequency vibration but do not operate correctly.*

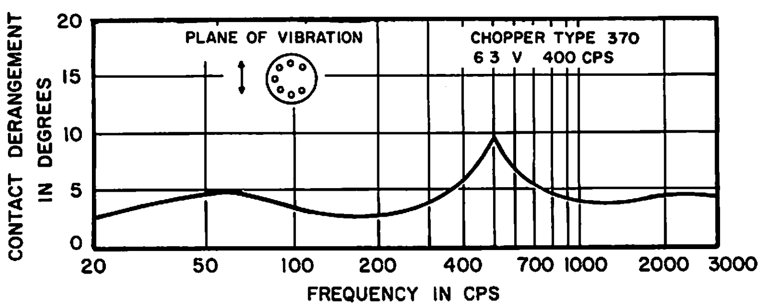


FIGURE 46. *Effect of 15 G vibration on a type 370 chopper.*

The drive coil and polarized magnetic circuit couple to the balanced armature at one end. The moving contact is mounted at the opposite end, well removed from the magnetic field to minimize stray pickup. Because unwanted vibrations act on the armature at its center of gyration, they produce no rotation, while the drive field acts at the end of the armature to develop maximum torque—leading to an efficient drive mechanism that requires low drive power.

Figure 46, shows that the effect of external vibration is isolated by this balanced structure. The curve, from a typical series 370 chopper, shows contact derangement as a function of frequency for vibrations of constant 15 G in the plane that is most sensitive to vibration. The natural resonance of the armature is seen to be about 500 CPS. Even here, however, external vibration has little effect on chopper action. The curve was obtained with the 400 CPS voltage to the drive coil at 5.7 RMS volts; that

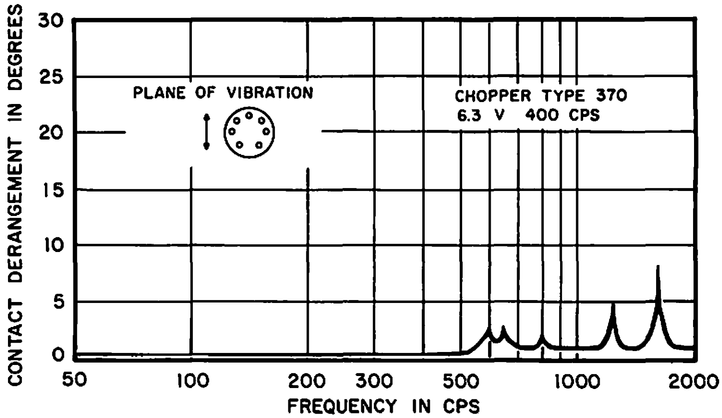


FIGURE 47. *Type 370 chopper vibrated in lesser plane.*

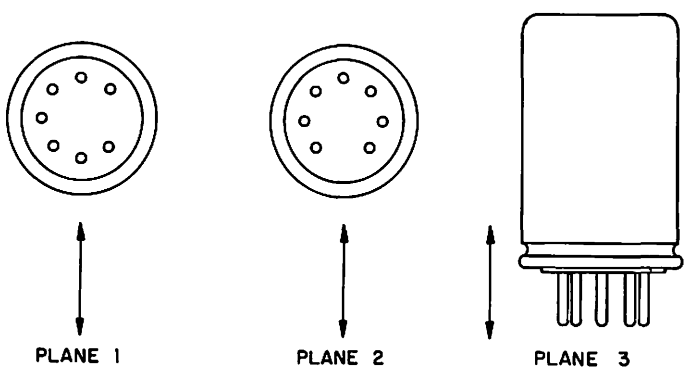


FIGURE 48. *Plane 1 delivers vibration parallel to the direction of contact motion.*

is, with the driving force at the lower limit of its specified tolerance. This curve shows what can be expected under the most adverse conditions.

Figure 46, examined the worst possible conditions for a type 370 chopper, using the plane of maximum contact derangement. Figure 47, is similar data, but taken at right angles to the direction of Figure 46, and its negligible derangement will be obvious. Most AIRPAX miniature choppers having a 7 pin base, whether with solder terminals or plug in, have

their internal works aligned with the illustration of Figure 48. Plane 1 is the direction of contact motion, plane 2 at right angles, plane 3 is of course parallel to the frame, and the chopper can.

LOW FREQUENCY CHOPPERS. The attack of a changing environment has been confined up to this point to 400 cycle choppers. At power frequencies of 50 or 60 cycles, components are often not asked to withstand the degree of change found in aircraft or missile equipment. However,

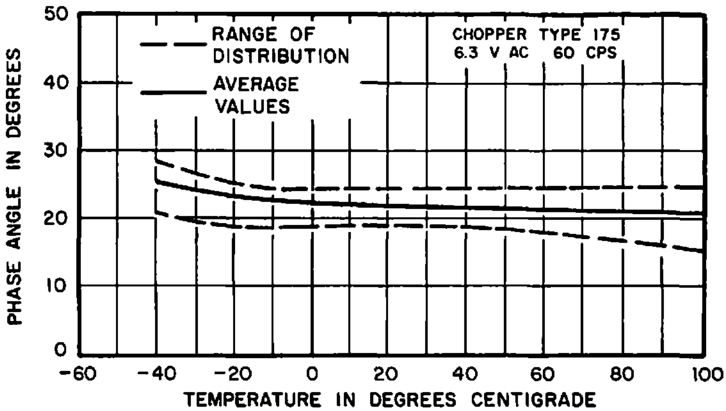


FIGURE 49. *Phase angle spread of a large group of type 175 choppers.*

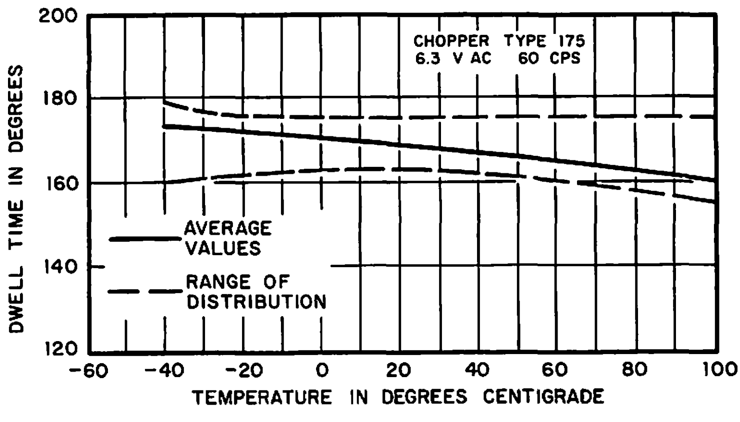


FIGURE 50. *Variation of dwell time with temperature and with production, type 175 choppers.*

the 60 cycle chopper can be built to withstand extremes as readily as at aircraft supply frequencies.

Figures 49 and 50, describe the limits of a family of curves taken on choppers type 175, when operated at 6.3 volts, 60 cycles, and subjected to an ambient varying from —40° to +100°C. It should perhaps be repeated that individual units do not fall in a random fashion within these boundaries—the upper and lower limits can be considered as describing two

individual choppers out of a large group from which this data was derived.

The effect of life up to 10,000 hours is shown in Figures 51 and 52. The units operated in "dry" circuit conditions, except for periodic recording of dwell time (every six hours with automatic equipment). During the several minutes on the recorder, 50 millivolts at 50 microamperes appears

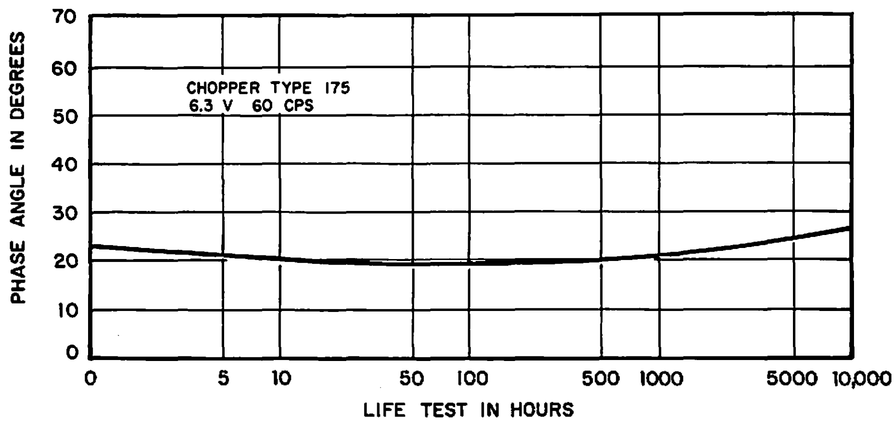


FIGURE 51. *Change of phase angle during 10,000 hours of operation, type 175 chopper.*

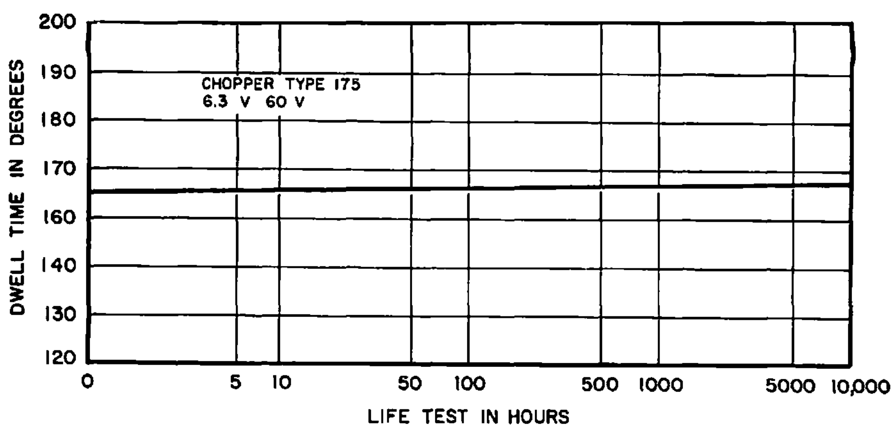


FIGURE 52. *Dwell time of a type 175 chopper operated for 10,000 hours.*

across the contacts, not enough to "clean up" the contact surface. As the choppers were operating satisfactorily at the conclusion of this particular test it is difficult to predict how long they might last—apparently they will continue indefinitely.

Figure 53, shows the amount of contact derangement from a vibration table excursion of 0.05 inches. Planes 1 and 2 of Figure 48 are plotted, being the worst and next worst planes. The contact derangement is usually symmetrical, appearing as a symmetrical modulation of the phase angle. Being symmetrical, it often does not cause offset in a DC amplifier, and as a 60 cycle chopper amplifier pass-band will be about 15 cycles, the vibration component may not appear in the output. In any case the chopper is undamaged. Just as shown earlier for the type 300, Figure 41, at high vibration frequencies the phase modulation becomes total, i.e., the chopper will follow the vibration.

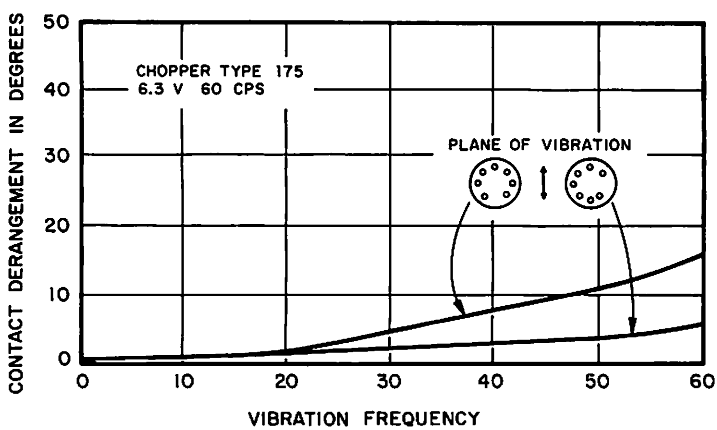
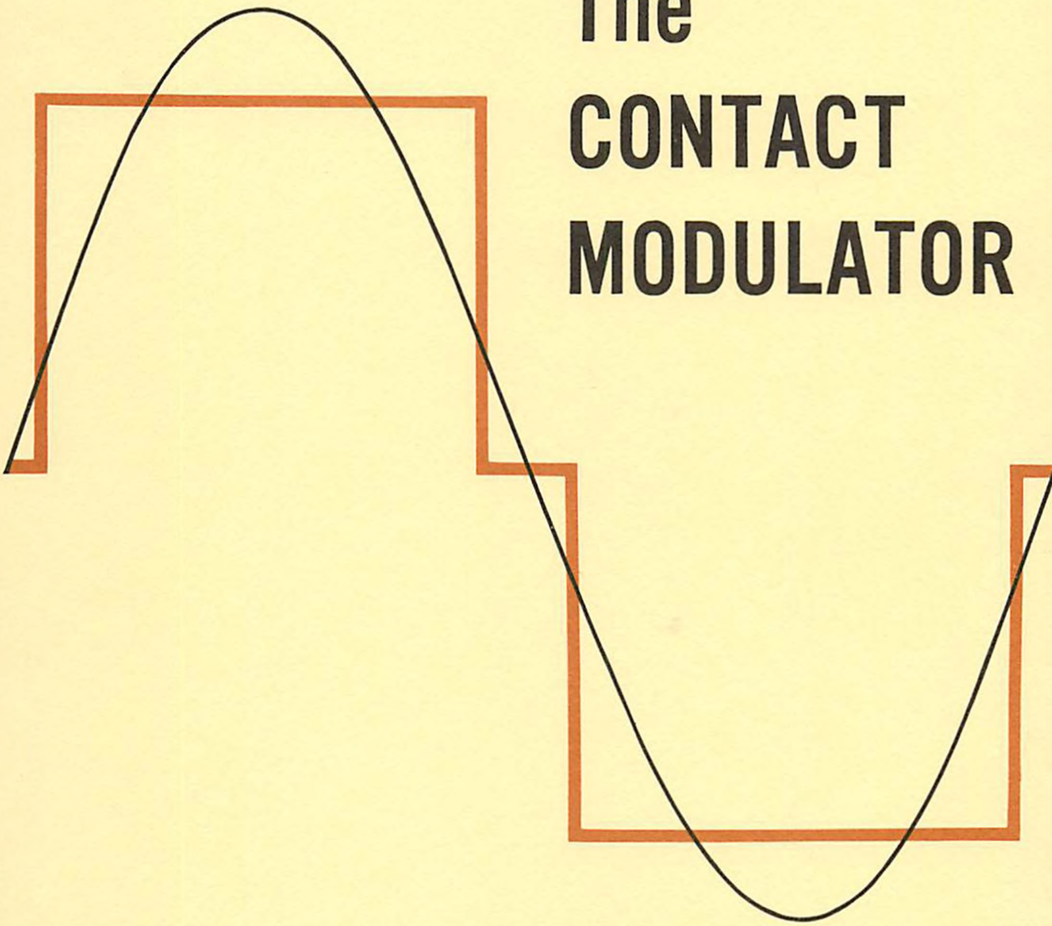


FIGURE 53. *Effect of vibration on a type 175 chopper.*

Types 300 and 370, at 400 cycles, and type 175, at 60 cycles, can only be good examples of characteristics observed on a few popular chopper models. Obviously other designs will have different response to ambient variations, and the user should ask for data on the model he finds most suitable for his application.

AIRPAX

The CONTACT MODULATOR



PART 5: Noise in Chopper Circuits

THE FIFTH OF A SERIES

AIRPAX

THE

CONTACT MODULATOR

PART V

NOISE
IN
CHOPPER CIRCUITS

WRITTEN BY

THE ENGINEERING STAFF

OF

AIRPAX ELECTRONICS

INCORPORATED

Cambridge, Maryland • Fort Lauderdale, Florida

H. A. COOK, President

DR. DAVID A. ROBINSON, Vice President, Engineering

FOREWORD

Choppers are frequently called upon to deal with minute values of direct current where the customary type of amplifiers require assistance. Noise originating in a chopper-amplifier system may erroneously be charged to the chopper.

Part V of the Contact Modulator series points out such noise sources and describes remedial actions. A section is also devoted to a comparison, by measurement, of choppers available on the market.

AIRPAX ELECTRONICS
INCORPORATED

TABLE OF CONTENTS

An assembly line at the Airpax Cambridge Plant

SECTION I
NOISE SOURCES

NOISE IN CONDUCTORS. Johnson noise, or random noise from electron motion due to thermal agitation, was reported by Johnson[1] in 1928. As shown by the expression $E_{eff} = \sqrt{4KTRF}$ (see PART II), the noise varies with temperature, T, resistance R and bandwidth F. Figure 1, is an oscilloscope observation of noise in a 100,000 ohm resistor over a bandwidth of 180 KC, as observed with the aid of a hushed transistor amplifier. From the equation, in which temperature T is given in degrees Kelvin, it can be surmised that cathode emission noise in vacuum tubes can be substantial. The Gaussian noise, white noise, random or Johnson noise, or some other name if you wish, tends to set the lower amplitude limit of signals which it is practical to observe. The value given in the equation is effective; peak values are about 8 times higher.

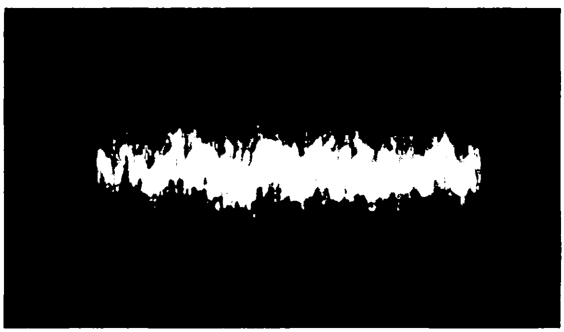


FIGURE 1. *Johnson noise has a uniform frequency distribution.*

Resistors also develop (when carrying current), a contact noise, at solder joints for example. Special low noise resistors may use welded connections to avoid this. Composition resistors are a molded assembly of carbon granules having many contacts, and may accordingly[2], in fact usually do, exhibit very substantial noise levels. These are often in the vicinity of 50 microvolts of noise per volt drop across the resistor concerned. Noise of this nature is negligible if the signal voltage across the input resistor is low, and becomes of greatest concern at high resistance values and moderate voltages. Considerably lower noise will be found on deposited carbon resistors, generally being in the vicinity of 0.5 microvolts per volt. Metal film types will exhibit about 0.2 microvolts per volt and are

[1] Thermal Agitation of Electricity in Conductors. J. B. Johnson, Physical Review, Vol. 32, July 1928, pages 97-109.

[2] Resistance Fluctuations in Carbon Microphones and Other Granular Resistances. C. J. Christensen and G. L. Pearson, Bell System Technical Journal, April 1936.

often used for high resistance circuits operating at low levels. The presence of a considerable amount of very low frequency noise has been observed [3] on deposited carbon resistors, enough to make them unusable for DC or low frequency amplifiers.

In theory, at least, wire-wound resistors should exhibit zero contact noise except where connection is made to the resistor. Unless non-inductively wound, these are susceptible to excessive hum pickup; the user must beware of curing one form of noise and introducing another. When moderately strong AC fields are present, such as may occur even from a chopper drive coil, special care is required, particularly at high resistances. The inductive cancellation of purportedly non-inductive resistors may not be sufficient. If sectionally wound the hum linkage may not be equal and magnetic as well as eddy current shielding of the resistor may be required.

NOISE IN THE CHOPPER. When chopper noise is specified (see PART II, Definitions), it refers to noise developed inside the chopper and

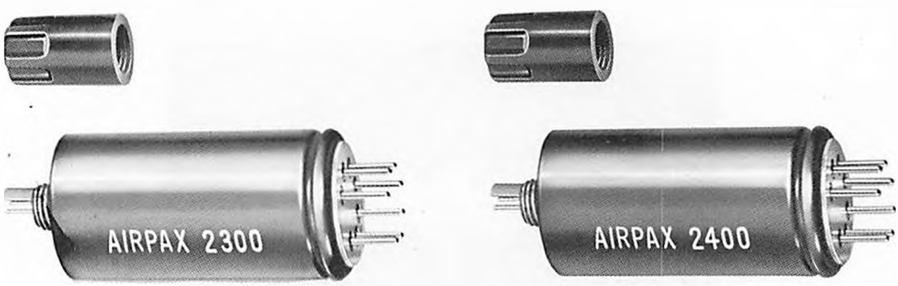


FIGURE 2. *Low noise choppers with drive coil leads brought to top terminals.*

present at the chopper contact connections. BUT! — *it is a fact that with presently available choppers of advanced design, the noise from the contacts is usually less than other circuit noise.* Choppers are now available having residual noise levels in the vicinity of 1 microvolt at impedance levels of about 1000 ohms. Airpax models 2300 and 2400 are examples, see Figure 2. The coil leads are brought out the top to make isolation easier for the user; actually, it is possible to use coil leads out the base without greatly affecting the noise, at least at moderately low impedances. The capacitive and resistive coupling between pins varies considerably with the type of socket used. Teflon and glazed ceramic sockets are usually satisfactory.

USE OF A DIFFERENT CARRIER. When strong power frequency fields are present, it may be helpful to drive the chopper from some other frequency. For this reason, we find 400 cycle choppers powered from a 60

[3] Operational Amplifier has Chopper Stabilization. David A. Robinson, Electronics, September 1956.

cycle line. Airpax type 202 tube inverter, or type 220 transistor inverter, can provide a suitable driving source. Filtering can then provide additional rejection of the power frequency, and the effect of stray fields reduced to a very low level.

CAPACITIVE PICKUP. Electrostatic and resistive noise coupling from drive to contacts is considerably reduced if the coil voltage is balanced to ground. This is nearly always helpful; it is of particular importance if the coil leads appear at the base with the contact leads.

In Figure 3, C_1 and C_2 are the stray coupling capacitors from either end of the coil which is referenced to ground by the center-tapped resistance R_1, R_2. This is redrawn in the more conventional bridge form in

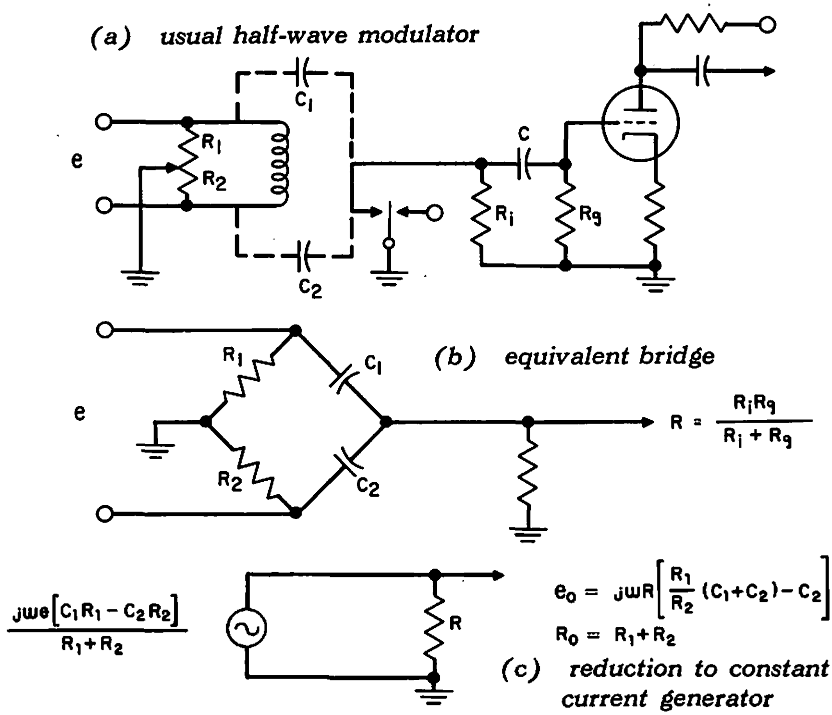


FIGURE 3. *First step in noise reduction is balanced drive voltage.*

Figure 3(b), with R_i and R_g lumped in parallel (assuming negligible source impedance). In practice, C_1 and C_2 are in the order of 10^{-15} farads so that

$$R_1, R_2 << \frac{1}{j\,\omega\,C_1}, \frac{1}{j\,\omega\,C_2}$$

This suggests using Norton's theorem to reduce the bridge to an equivalent constant current source. It is now possible to write the expression for e_o, the noise appearing on the grid as given in Figure 3(c). Since $R_1 + R_2 = R_o$, the constant impedance value, it is convenient to eliminate R_2 and choose

R_1/R_0 as the independent variable. We see that e_0 is linear in R_1/R_0 and proportional to $j\omega eR$, with bridge balance at $R_1C_1 = R_2C_2$.

A noise level due entirely to electrostatic coupling is shown in Figure 4, representing about 100 microvolts rms. Of course, the terminology "rms" is not valid; in this and succeeding scope pictures, the level depicted is measured by substitution of a known AC level. Accordingly, it is much more correct to say the peak-to-peak value shown is about 280 microvolts. We will be obstinate and continue to use rms although in error.

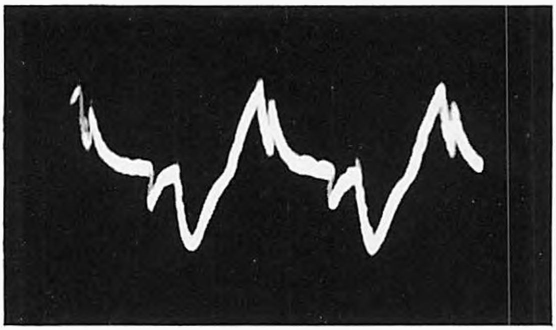
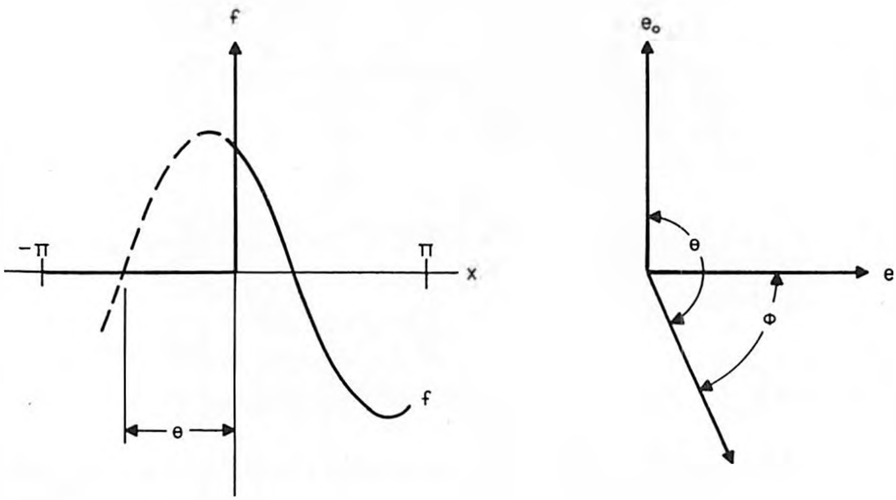


FIGURE 4. *Capacitive coupling, drive voltage to contacts.*



(a)

(b)

$f(x) = 0 \qquad x \in [-\pi, 0)$

$f(x) = e_o \sin(x - \theta), \; x \in [0, \pi)$

e = COIL VOLTAGE
e_o = PICKUP TO GRID
x = CHOPPER CONTACTS
Φ = CHOPPER PHASE ANGLE

FIGURE 5. *Electrostatic noise.*

A noise component will be introduced if leakage exists between the coil terminals and the contacts. A grounded shield is a good solution, as well as good quality sockets. The importance of this is obvious if the insulation resistance is considered. If the circuit impedance is one megohm, 6 volts is applied across the socket to ground in an unbalanced fashion, and the insulation resistance is as high as 10,000 megohms, the noise level would be 600 microvolts rms.

It is instructive to consider the effects of this electrostatic noise on the output. The noise is sinusoidal and operative during the half cycle the grid is ungrounded. Figure 5(a), shows this waveform and describes it analytically where θ is the angle by which e_0 leads the contacts. For a DC output, we assume perfect clamping so that the input DC content appears intact after demodulation. For AC output, we are interested in the fundamental

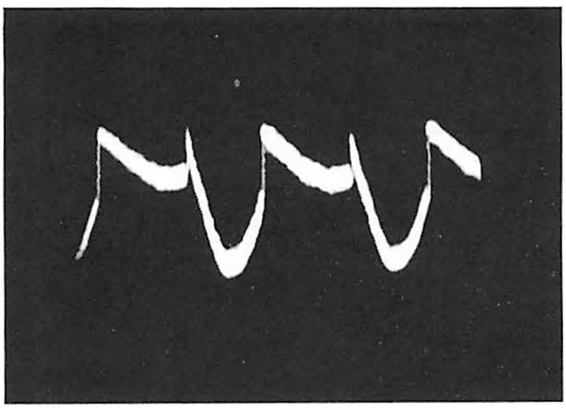


FIGURE 6. *Approximately 170 microvolts of electromagnetic noise pickup.*

of this waveform. Expanding in a Fourier Series we find:

$$f = \frac{e_0}{\pi} \cos\theta + \frac{e_0}{2} \sin(x - \theta) + \ldots$$

The appropriate term can be selected depending on the nature of the desired output. Figure 5(b), shows the phase relationships for a chopper with a phase angle of 65°. The j operator in the expression for e_0 indicates quadrature with e so that $\theta = 90° + \phi$. Observation of the series indicates the desirability of a zero phase angle chopper since at $\theta = 90°$, the DC level vanishes and the fundamental is in quadrature with the contacts and would, for example, contribute no torque to a two-phase motor load.

MAGNETICALLY INDUCED NOISE. Electromagnetic noise may also be found within the chopper, as illustrated in Figure 6, equivalent to a signal of about 170 microvolts rms. Such a noise pattern, if it were within

the chopper, might be originated as shown in Figure 7. During the half cycle the grid is grounded, magnetic lines of force intercept the loop so formed and generate a voltage across R_g as shown in Figure 7(a). Note, however, that this loop can also lie outside the chopper. It is wise to return the chopper ground lead directly to the point where the first stage cathode bias resistor is grounded, running it close to the grid lead to keep the area of the loop small and free from external magnetic fields. Distant chassis groundings are not advisable if noise is critical. Within the chopper case, the magnetic field from the coil intercepts this loop and generates a sinusoidal noise e_m of fundamental frequency as shown in Figure 7(b). The magnitude of this noise may be much less than 1 microvolt, or up to 100 micro-

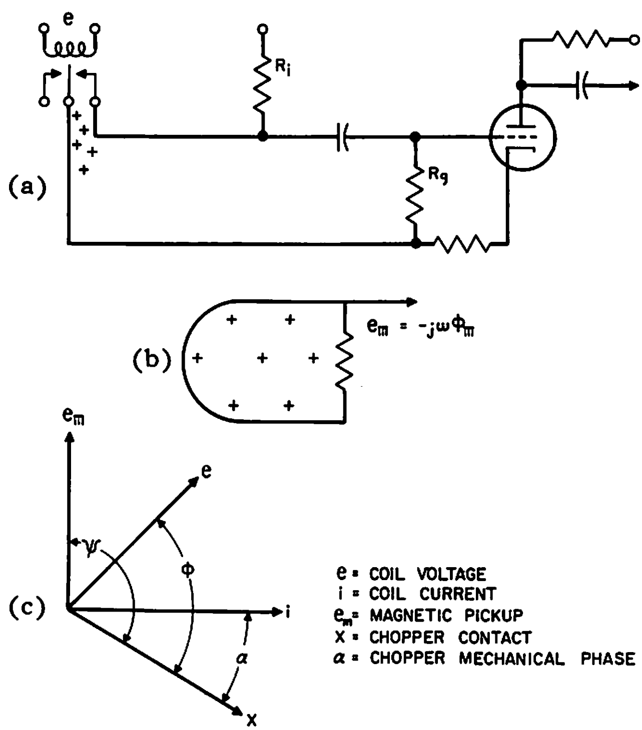


FIGURE 7. *Electromagnetic noise from within a chopper.*

volts peak-to-peak, depending on the chopper's internal arrangement and magnetic shielding. Referring again to the loop of Figure 7(a), one possible source of trouble can be eddy currents induced in the ground return line, from the chopper coil to the chassis itself. This noise cannot be balanced out as easily as the electrostatic pickup, but various means could be devised whereby a single turn from a conductor carrying the coil current and the offending loop can be juxtaposed to cancel the total loop flux. Obviously, this method is neither consistent nor too practical. This noise is sinusoidal

and occurs during the half cycle that the grid is grounded so that Figure 5(a), and the previous Fourier analysis again apply except that ϕ is replaced by ψ, the angle between the electromagnetic noise e_m and the contacts. As pointed out in Figure 7(c), the magnetic noise e_m is in quadrature with the coil current, and is dependent on α, the mechanical phase lag of the chopper. For the AC output case, we write the fundamental term as:

$$\frac{e_o}{2} \cos \theta \sin \psi - \frac{e_o}{2} \sin \theta \cos \psi$$

The in-phase component, just as the DC level, depends on θ (or ψ). But:

$$\cos \theta = \cos (90^\circ + \phi) = - \sin \phi$$
$$\cos \psi = \cos (90^\circ + \psi) = - \sin \psi$$

and it now becomes apparent that if we set

$$e_o = e_m \frac{\sin \psi}{\sin \phi}$$

we could cause the electrostatic and electromagnetic noise to cancel.

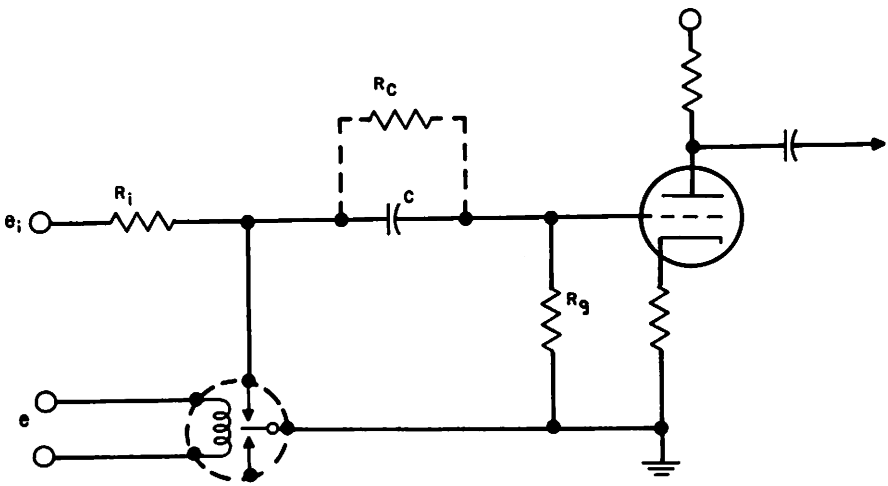


FIGURE 8. *Capacitor leakage can cause noise.*

Since choppers having noise levels of a few microvolts are available, it is probably best to use them. Hum cancellation is an old expedient; if necessary to reduce circuit noise it is often helpful, but variation in components may prove frustrating in production. Having eliminated trouble within the chopper, however, we must be as careful at the socket, and beyond the socket, to avoid the many sources of noise pickup.

EFFECT OF LEAKAGE RESISTANCE. Imagine in Figure 8, that C has a leakage resistance R_c across it. The chopper is periodically shunting

R_c across R_g. If the tube is drawing grid current, the grid will appear as a constant current source. If, say, this current is 1 microamp, R_g is 1 megohm and R_c is 100 megohms, a little calculation shows that the voltage across R_g varies between 1.0000 and .9990 volts as it is successively shunted by R_c. This is equivalent to 1 mv of noise at the input. There are two apparent solutions to this; operate the tube with cathode biasing at the grid current cross-over point (—.85 V DC for a 12AX7) to limit the current (say, less than .1 microamp) and use a high Q capacitor [4] such as mica that is clean and free of body leakage paths. Now, if I_g is .1 microamp and R_c is 1 tera-ohm (10^{12}), then the equivalent noise is reduced to only .1 microvolt. For this reason, self-biasing grid leak resistors such as are used in starved amplifiers, are not recommended.

Grid current can cause leakage, with resulting noise, similar to that shown in Figure 9. The presence of DC leakage paths of any kind must be avoided, and the common practice of assembling all components on ter-

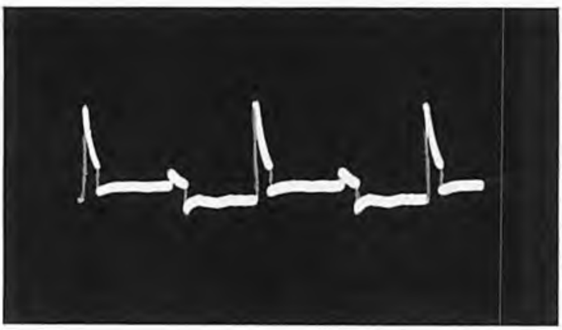


FIGURE 9. *Noise pattern caused by grid current.*

minal strips can easily cause trouble. In tube circuits, 300 volts DC is frequently used; if the amplifier responds to microvolts, the insulation resistance has to approach perfection unless isolation is provided. It is important to remember that the chopper will interrupt any voltage given it, and with unfortunate frequency this turns out to be things other than the signal under examination. A practical solution is to group sensitive components on separate terminal strips, or in some other way to interpose a metallic ground between high voltage DC and sensitive points.

Capacitors at times exhibit the ability to retain a small DC charge, sometimes even after a short circuit, for substantial time periods. Polystyrene dielectrics seem to be relatively free from this effect. The chopper circuits, of course, consider such a retained potential as signal, and offset appears.

[4] DC Amplifier with Reduced Zero Off-set. McAdam, Tarpley and Williams, Jr., Electronics, August 1951.

THERMAL JUNCTIONS. Dissimilar metals, solder joints, and thermal junctions of many forms can present little batteries in series with the chopper contacts. In Figure 10, there appears about 30 microvolts rms of noise caused by such a junction in a tube socket. Figure 11, shows the same sort of unwanted junction, but with in-phase hum pickup via a ground loop, to give a substantial total noise of about 100 microvolts.

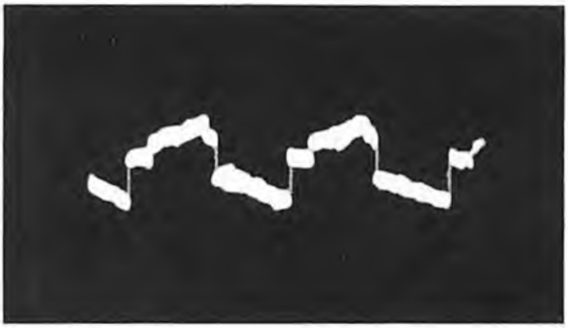


FIGURE 10. *Thermal junctions cause noise in chopper circuits.*

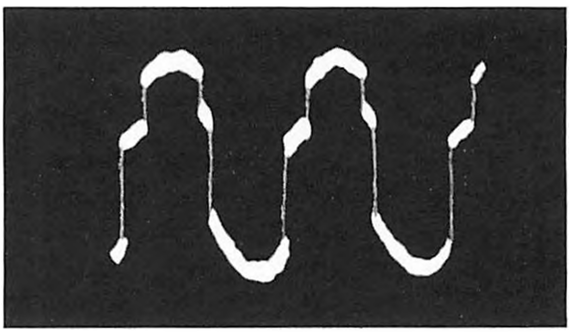


FIGURE 11. *Thermal junction plus ground loop hum.*

It will be observed that thermal junctions often appear when the ambient is first altered, and that they will return to zero, or near zero, when both parts of the junction resume the same temperature. This is distinctly different from a thermocouple, which delivers a voltage proportional to temperature with both of the dissimilar conductors at the same temperature at the point of junction. A frequent offender, delivering as high as several hundred microvolts of output with only a few degrees temperature differential, is the kovar to copper junction frequently found when hermetic seals are used. An ordinary hair drier can be used to locate the offending junctions; simply direct the hot air blast at the suspected spot for a few seconds and note the results.

The best cure is not to join dissimilar metals; failing that, to join conductors having low thermal junction potentials, using low thermal junction solder as described later. A much less desirable solution is to reverse two junctions in the hope they will cancel each other.

Joints are sometimes very difficult to weld or braze, particularly in the vicinity of delicate parts, and solder has been found [5] to develop substantial potentials when used to form joints. A "low noise" solder is recommended by solder manufacturers, having a composition of 70% cadmium, 30% tin. This has a wide plastic range of 360 to 470° F. (The usual radio solder 60/40, has a range of 362 to 374° F.). This cadmium tin solder flows rather well, but requires a flux more acidic than 60/40. A rosin flux, to which a small amount of acid has been added, seems satisfactory. The precise composition of such a solder was reported by one supplier to be 70.44% of commercially pure cadmium and 29.56% of tin, c.p. When used with copper wire the junction potential is said to be about 1/5 that of usual 50/50 tin and lead solders.

MICROPHONICS. Any chopper will transmit, to some degree, its mechanical vibrations to the surrounding supporting members, usually a chassis which will in turn transmit them to the tubes of low level stages. The obvious methods are used to combat this, such as the use of ruggedized tubes, preferably triodes. If at all possible, do not mount the input stage directly adjacent to the chopper. Place both chopper and input stage near the edge or corner of a chassis for stiffening and avoid very thin metal for the chassis. Mechanical standing waves could develop in a non-rigid chassis. The chassis itself, especially if fairly small in mass, such as a strip chassis, should be firmly mounted to the more massive sections of the apparatus. Rubber grommets as vibration insulators should be used with caution as they can often make the situation worse. Fortunately, microphonics are seldom a serious problem, and with a few precautions rarely cause noticeable offset.

[5] Contact Modulated Amplifier to Replace Sensitive Suspension Galvanometers. Liston, Quinn, Sargeant and Scott, Review of Scientific Instruments, Vol. 17, May 1946, pages 194-8.

SECTION II

REDUCTION OF NOISE

HUM IN INPUT TRANSFORMERS. The transformer input of Figure 12, permits almost complete input isolation, a high order of common mode rejection, and often considerably greater circuit gain. The transformer, however, avoids only partially the ground loop problem, so far as noise pickup is concerned, and inconveniently provides a coupling place for the field radiated from the chopper coil. Figure 13, shows the reasonably low figure of 25 microvolts rms of hum pickup from the chopper induced in the core of the input transformer.

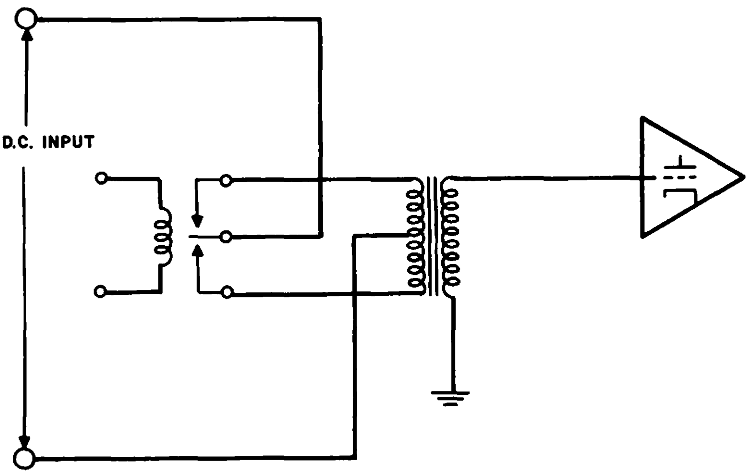


FIGURE 12. *Transformer input has many advantages.*

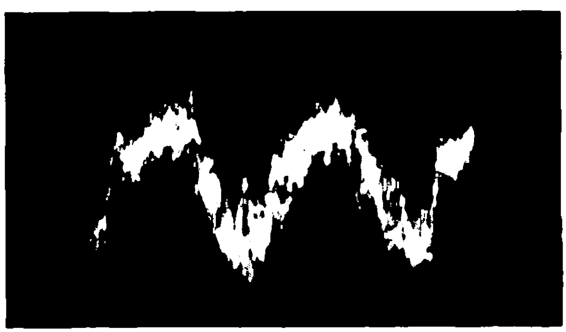


FIGURE 13. *Hum pickup in input transformers.*

Hum in input transformers is reduced by classical methods, such as the use of hum bucking construction as in Figure 14. In this so-called "core" type of lamination most of the external magnetic flux passes through both core legs inducing equal and opposite voltages in the two coils; hence, much of the noise (hum) is cancelled out. The Triad Transformer Corporation of Los Angeles produces several chopper transformers capable of excellent performance. One type is their G-24, shown in Figure 15, which uses hum bucking construction, a series of nested hum shields and leads

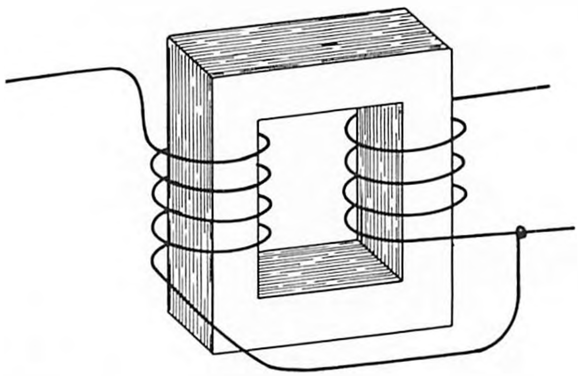


FIGURE 14. *Hum bucking construction reduces noise pickup.*

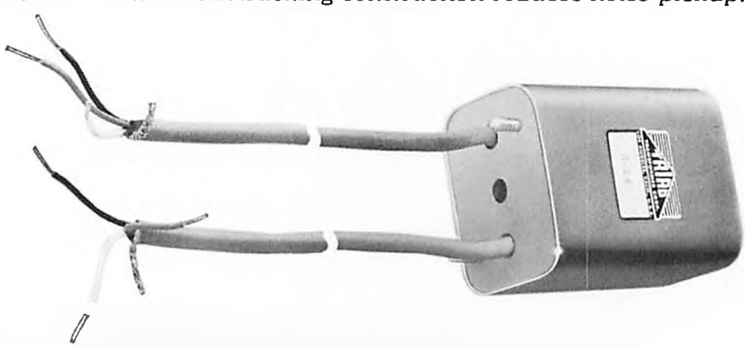


FIGURE 15. *Triad type G-24 has low hum pickup. Continuation of coil winding to output leads avoids thermal junctions.*

brought out directly from the conductors. This last apparent retrogression in the transformer art helps avoid thermal junctions from dissimilar metals and solder joints. Additionally, type G-24 has independent shields around primary and secondary, permitting ground connections to avoid ground loops, often a serious cause of noise. The primary to secondary capacity of less than 0.1 mmfd permits better balance at the input terminals and improved common mode rejection. Internal connections are silver brazed, a method apparently less subject to junction voltages than soldering.

The illustrated transformer, which has a 1:1 ratio, was set up in the circuit of Figure 12, but operating into a transistor input which terminated the secondary with about 10,000 ohms. Under these conditions, with an Airpax low noise chopper (type 2300) operating directly adjacent to the transformer, DC signal levels of 1 microvolt could be distinguished from the circuit noise. Apparently the transformer is very well shielded, Figure 16. No measurement of common mode rejection was made; however, it is suspected that attenuation to a 60 cycle frequency could be very good.

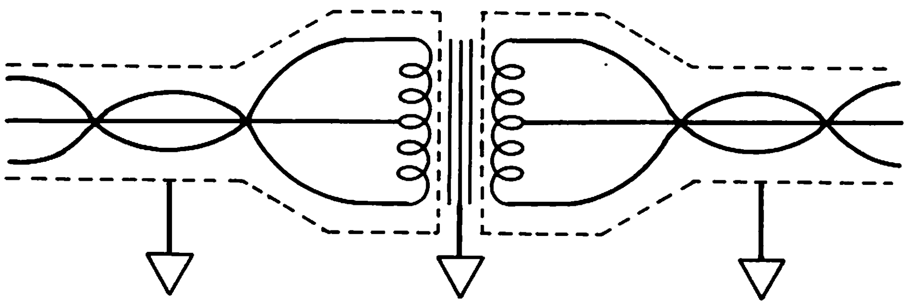


FIGURE 16. *Completely shielded windings of Triad G-24 permits improved balance and common mode rejection.*

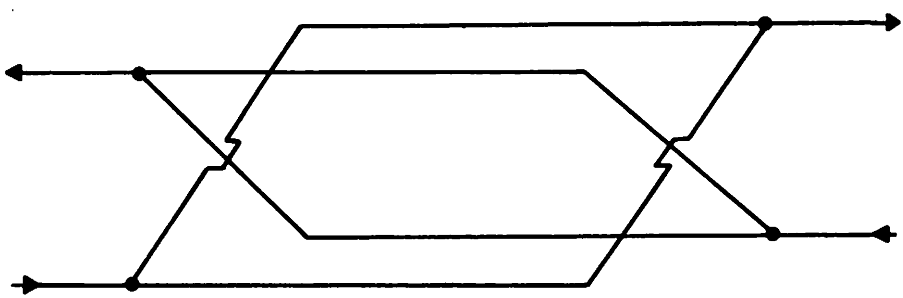


FIGURE 17. *Hum pickup can be reduced on long lines by multiple conductor transposition.*

LONG LINES. Long runs from thermocouples and other transducers frequently become necessary, and means must be provided to reduce hum pickup. Obviously, one way is to twist the pair, in the hope that cancellation will reduce the pickup. An extension of this thinking [6] is shown by the interconnection of four conductors as in Figure 17. Conductors which are joined should be diagonally opposite, as illustrated. This method is reported to give substantially less hum pickup compared to the twisted pair.

[6] Radio Engineers' Handbook, F. E. Terman, page 174.

SPIKE NOISE. In Figure 18(a), a metallic body, A, is moving away from a positively charged body. The electrostatic field readjusts itself in time so that fewer lines terminate on A, but go instead to ground. This requires positive charges to flow from ground onto A to neutralize the decreasing electrostatically induced negative charge. The result is a voltage induced across R due to the mechanical motion of a conductor in an electrostatic field. This seemingly obvious fact can now be extended to explain a large number of noise phenomena where mechanical motion is involved. The source of the electrostatic field is not difficult to understand since mechanical motion will induce frictional static electricity in a large number of dielectric materials. It should also be noted in Figure 18, that relative motion of any one body (A, Q, or ground) with respect to the other two is sufficient.

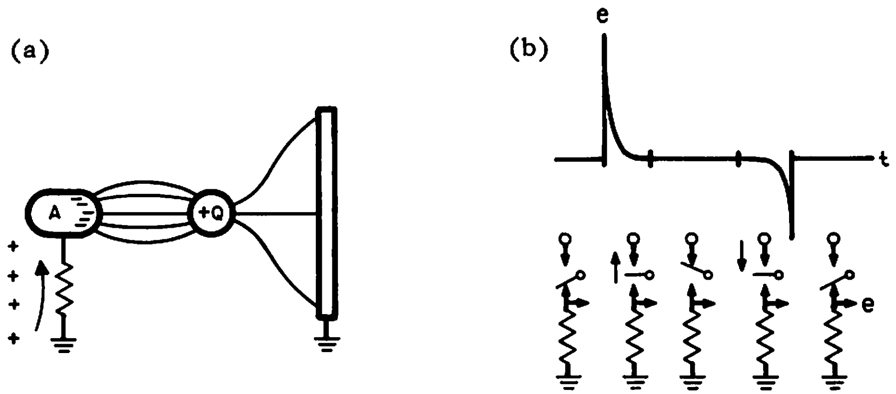


FIGURE 18. *Noise can be caused by insulating materials.*

For example, an insulated lead wire vibrating against a ground plate will generate noise in the wire; the insulation supplying the static electricity and the wire and insulation moving relative to the ground, supplying the needed mechanical motion. In fact, the basis of cable noise lies here. One can see from this that care must be exercised in choosing insulating materials for lead wire jackets, spacers and shock mounting. Glass, teflon and silicon are a few insulators that show surprising ability to pick up and retain large amounts of frictionally induced static electricity.

Spike noise is occasionally observed in some choppers and is characterized by a pulse of noise occurring just as the contacts break. This pulse has a very sharp rise time followed by an exponential decay as shown in Figure 18(b). Investigation has satisfactorily shown this to be merely another form of static field noise. A microscopic quantity of insulating material between the contacts themselves can supply the minute but sufficient static field, while the contacts, in the act of separating, supply the necessary relative motion. The spikes often come in pairs of opposite

polarity (one for closing contacts, the other for opening) and contain little DC level and only high harmonic AC content. Such spike noise often may appear and disappear completely over a few moments time. Fortunately this noise also tends to disappear with time and working of the contact.

Static field noise can appear in cables,[7] from mechanical vibration, although commercially available cable is largely free of this trouble. Microdot supplies a low noise coaxial cable in which the insulation is reported to be impregnated with a metallic powder. Such noise has also been observed in the insulation of wires connected to choppers. Because of the low energy content and the high harmonic characteristic, this type of noise can easily be present and not be observed at all. The usual chopper stabilized amplifier employs considerable input and output filtering of the chopper amplifier loop, and frequently the response is as low as 1 or 2 cycles.

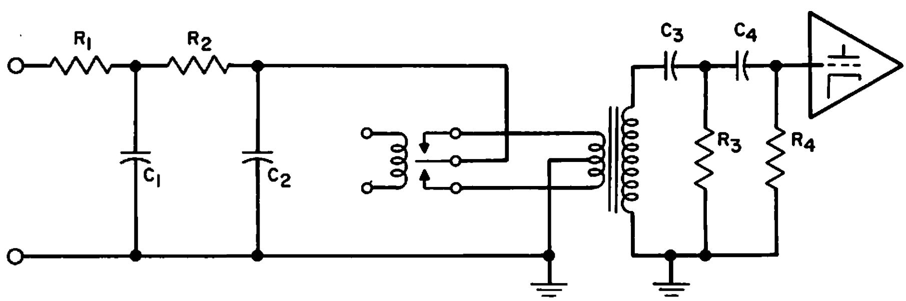


FIGURE 19. *Filter networks reduce noise problem.*

FILTERING OF NOISE. Low frequency disturbances, such as slowly changing grid currents, can result in noise, as they may be chopped and subsequently amplified.[8, 9] In Figure 19, the input grid is separated from the transformer secondary by a high pass filter which need only pass the carrier frequency as its lower limit.

CAPACITIVE DISTURBANCE. Low frequency disturbances, such as shock, or merely the operator's hand, can cause capacitive changes resulting in noise; of course this is at its worst in high impedance circuits. It may prove necessary to shield input leads and terminals. Sometimes it is necessary to arrange for special wire support, or arrange leads to have a very low capacity to ground, to avoid shock and vibration disturbance.

[7] National Bureau of Standards Technical Report, No. 1645, Dr. T. A. Perls.

[8] DC Amplifier Stablized for Zero and Gain. Williams, Tarpley and Clark, AIEE Transactions, Vol. 67, 1948.

[9] R. E. Tarpley Patent 2622192, Measuring System with Grid Current Suppression.

GROUND LOOPS. It is frequently difficult, particularly in high gain stabilized amplifiers responding to 60 cycles, to avoid hum loops caused by common ground impedances. Most amplifiers require a ground return; when they do, as in Figure 20, they are confronted with inconveniently high noise. An attempt is often made to group ground connections to render them harmless to each other, so we have the one point ground theory and the sub-grouping theory, as in Figure 21 (a) and (b). It may prove helpful to employ a ground strip, or arrange components so that hum producing connections are closer to earth than the sensitive input circuits, as in Figure 21(c), where point X might be an input circuit and Z, a later stage.

If the components under examination are tied directly to a large metal surface and sufficiently spread about, the hum pickup due to ground loops will probably disappear completely. We have brought the ground connection directly up to the parts involved and have spread out the common

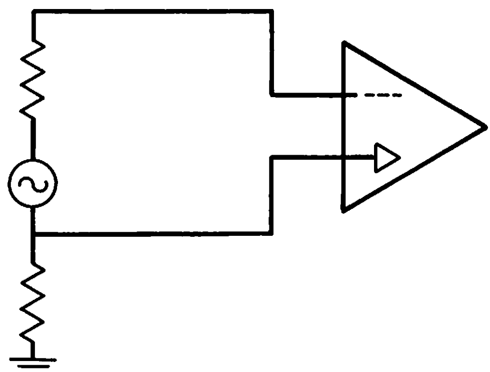
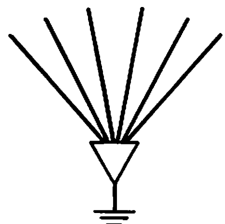


FIGURE 20. *Ground loops are difficult to eliminate.*

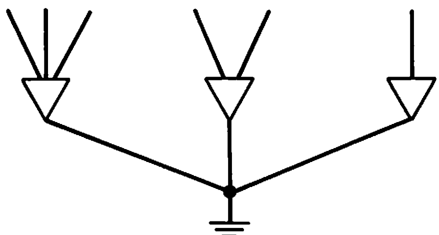
impedances so that common coupling is reduced to a vanishing point. It may not even prove necessary to physically ground the metal sheet. However, metal sheets several feet wide do not prove too portable, and attempts must be made to avoid hum loops. Hum need not be merely common impedances; often hum is induced directly into the amplifier chassis, from a chopper, or nearby power transformers or wiring. The ensuing eddy currents may then appear across a section of the chassis and effectively be in series with the sensitive input circuit. Sometimes the substitution of brass or aluminum for steel, in a chassis, can avoid hum, and sometimes it can increase it, by reducing the shielding of parts. Shielding has been observed to increase hum by increasing the coupling to the input circuit, as can readily happen if hum voltages are induced in a ground loop of the input.

COMMON MODE REJECTION. Common mode rejection refers to the attenuation presented to stray hum noise by balanced input circuits

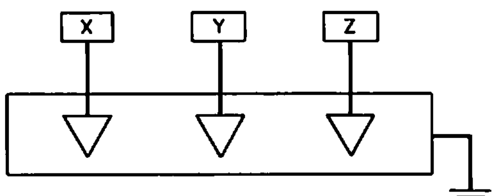
such as are provided by chopper-transformer combinations, as in Figure 22. The problem arises when long runs are made, as to thermocouples and strain gages, when it may become necessary to rescue the signal from hum which has been reported as high as several hundred millivolts.



(a) some follow the single point theory



(b) others favor groups of sub grounds



(c) a "line-up" sequence can also be used

FIGURE 21. *Various grounding systems.*

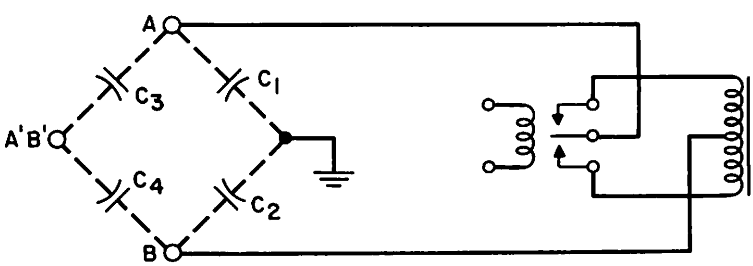


FIGURE 22. *Hum pickup is reduced by balanced input.*

Another problem is that it is often easiest or best to ground the transducer; for example, a thermocouple may be welded or soldered directly to parts of a jet engine. This demands a completely floating input, a difficult requirement for anything other than a mechanical chopper combined with a transformer.

In Figure 22, the values of C_3C_4, the capacity to the power line, are assumed equal. If C_1C_2 are equal, the bridge will be balanced. In testing for common mode rejection, it is customary to apply 60 cycles at perhaps 10 volts between point A^1B^1 to ground, using equal coupling capacitors C_3C_4, and measure the 60 cycle component discovered at the amplifier output. It will be observed that the chopper should be isolated from the drive coil and that the symmetrical balance of the input transformer is important.

MEASURING NOISE. We have found it necessary to begin with a large sheet of metal, joined directly to earth with a reasonably heavy conductor. It is helpful, but not necessary, to also have a shielded room particularly if strong radio frequency fields are present. A suitable size is a sheet of galvanized steel covering the entire table top, with a No. 10 copper connection to an excellent ground. At 1 microvolt these precautions will be found necessary. The various items of equipment will, as a rule, perform best when grounded directly where they sit, with a good solder joint.

Amplifiers have been described [10] with noise levels in the microvolt region; obviously the amplifier has to be good to measure chopper noise levels lying in the region of random noise. It is also important that the measuring voltmeter be considered.

For laboratory purposes, it is practical to calibrate an oscilloscope screen in terms of the output of a known voltage and a calibrated attenuator. This provides a peak-to-peak reading with an rms calibration, and the resultant number may bear little or no relation to offset. If energy content is wanted, as will be true if offset is to be considered, the noise output should be read on a thermocouple meter for true rms value, or rectified and read as DC, yielding an average value.

In any case, the specification of noise requires the complete detailing of impedance, amplifier frequency response, and of the measuring meter if a usable answer is to be obtained.

[10] Hushed Transistor Amplifiers, by W. K. Volkers and N. E. Pedersen; a paper presented at the National Electronics Conference, October 1955.

SECTION III

COMPARATIVE MEASUREMENTS

SOME NOISE FIGURES. A considerable group of Airpax choppers were tested in company with the choppers of 3 competitors, whom we shall term A, B and C. They reveal that Airpax is not alone in noise reduction, in fact B does quite well and builds a good chopper which exhibits excellent uniformity.

The data displayed was taken with the circuit of Figure 23. The Airpax input amplifier shown is similar to the preamplifier previously described (Part II), and is used when measurement is undertaken at im-

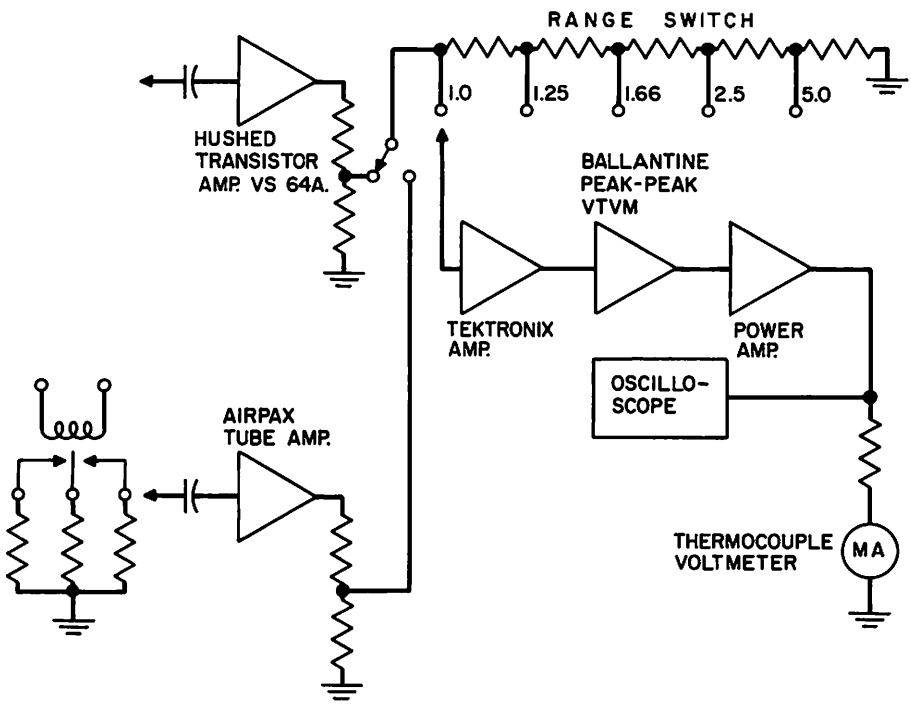
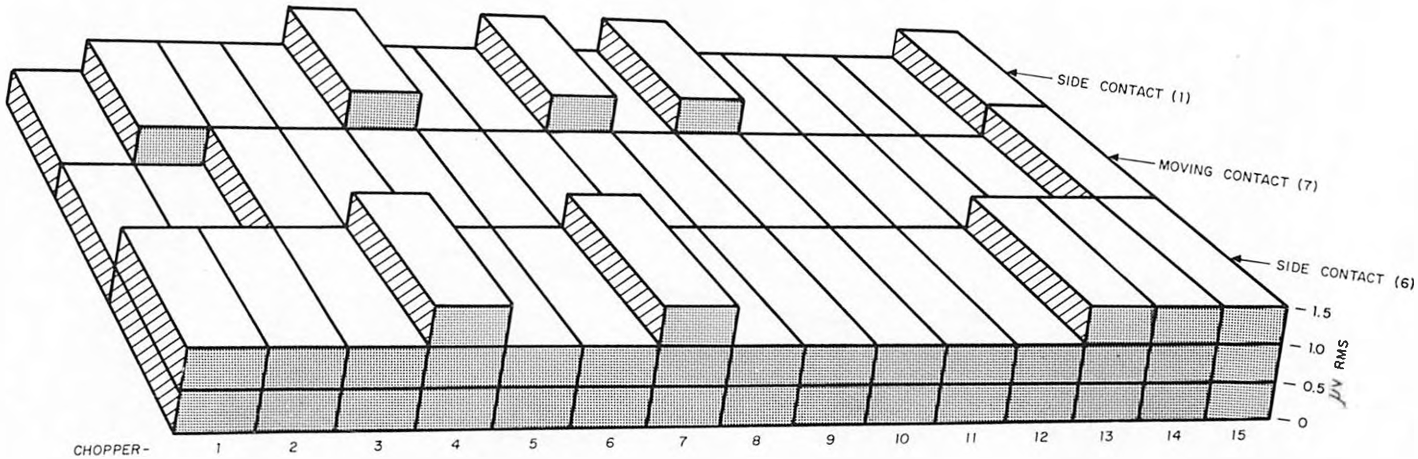


FIGURE 23. *Special amplifier arrangement permits both peak-to-peak and rms readings, also oscilloscope observation.*

pedances too high for the transistor amplifier. Type VS64A has 10,000 and 100,000 ohm impedances. Multiplier switch resistors are selected to permit easy reading of the square law scale of the thermocouple meter. Noise data shown subsequently is as read in rms on the thermocouple milliameter (in volts). A good capacitor is used ahead of the VS64A. The input electrolytic capacitor present in the amplifier exhibits substantial leakage initially, which appears as chopper noise. Several minutes of operation are required



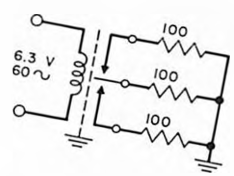


FIGURE 24. *Measured noise levels of fifteen Airpax type 2400 choppers.*

before the leakage reduces to a sufficiently low value. The system noise at the lower input impedances (from 100 to 10,000 ohms), was below 0.5 microvolts, increasing to about 2 microvolts at 100,000 ohms.

All the charts, shown in 3 dimensions in an attempt to present a clear and complete picture, are measured at 100 ohms circuit impedance, per Figure 23, and with the drive coil voltage center-tapped and grounded. Precautions are taken to insure against extraneous noise pickup in order that noise due to chopper action alone may be evaluated. Test conditions are of course identical for all choppers under consideration. In each case the system noise is checked and deducted from the output indication in order to provide a true measurement of chopper noise.

Figure 24, shows a sampling of 15 Airpax choppers, and can be compared with Figure 25, choppers as supplied by manufacturer B. These did not exceed 1.0 microvolt; had we 15 or more, instead of 6 samples, we might have found some noisier units. As it is, Airpax choppers do not exceed 1.5 microvolts, and since the system noise is 0.5 microvolt, it does seem that we might claim there is very little difference. The choppers of manufacturers

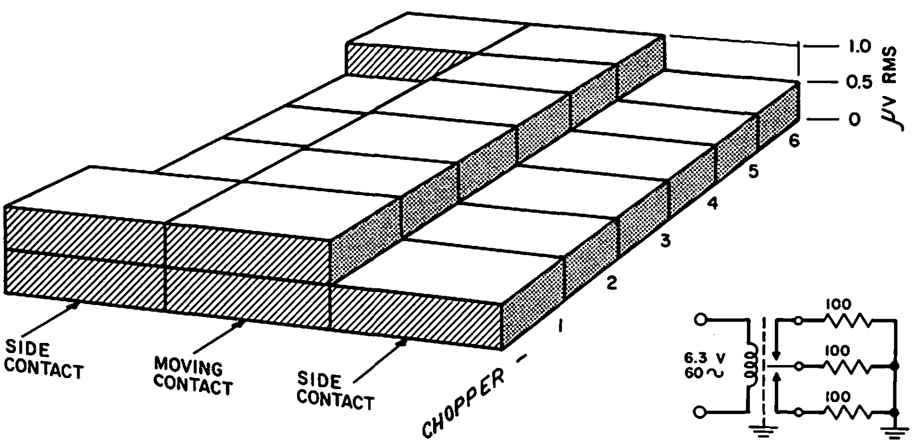


FIGURE 25. *Measured noise levels of six choppers, manufacturer B.*

A and C, shown in Figures 26 and 27, did not do nearly so well. (But it should be quickly noted that 6 to 14 microvolts is really not much noise. How good is your amplifier)?

The circuit impedance and noise relationship merited some study. In one chopper, a very small portion of the electrostatic shield between coil and the moving contact was left out, and the resultant plotted as the solid

line in Figure 28. The dotted line is a normal chopper. Incidentally, choppers from B and from Airpax both gave results almost identical with the dotted line. Apparently electrostatic coupling has very negligible effect below 1,000 ohms, insofar as noise from the chopper itself is concerned.

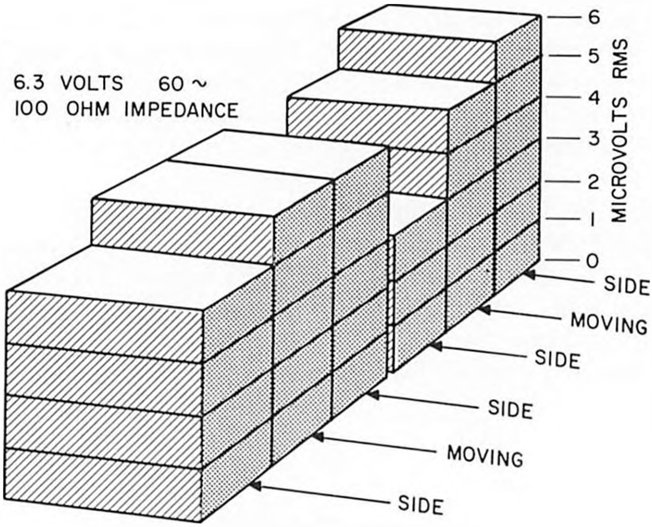


FIGURE 26. *Measured noise levels of two low noise choppers, manufacturer A.*

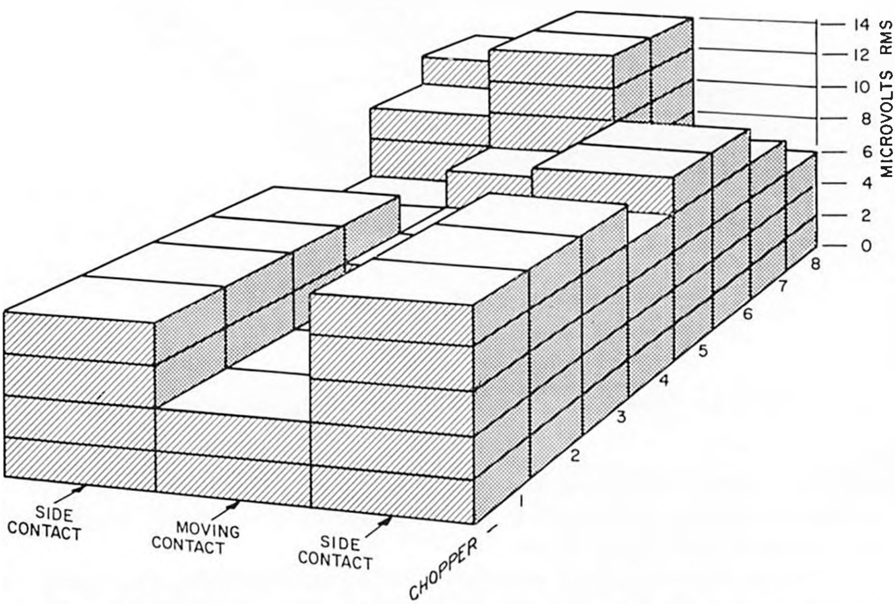


FIGURE 27. *Measured noise levels of eight choppers, manufacturer C.*

Figures 29 and 30 again compare Airpax and B, but here Airpax has a slight edge, and perhaps we did not test enough Airpax units. The efforts of manufacturers A and C showed noise values between 8 and 80 microvolts and were not plotted.

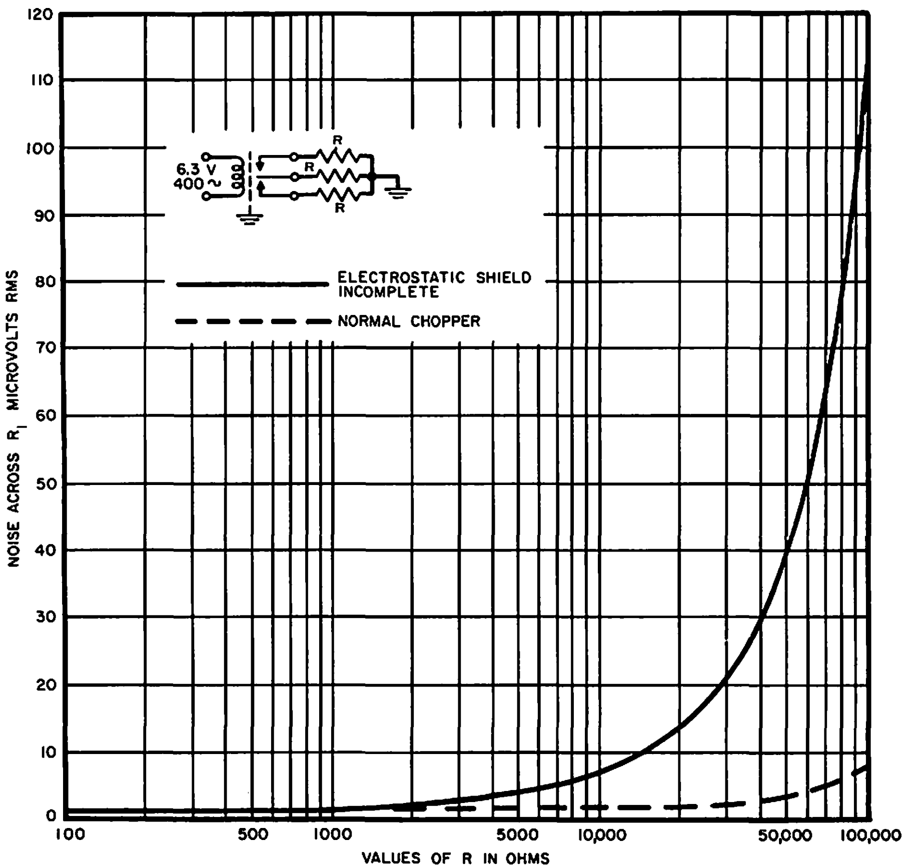


FIGURE 28. *Effect of circuit impedance on chopper noise.*

SOURCES OF NOISE. As has been observed, there are many sources of noise in chopper amplifiers, and the chopper is only one possibility. Since its characteristics are known and predictable, the probable effect of chopper noise can be decided when the amplifier gain is known.

Airpax noise specifications are usually written to include all possibilities, i.e., they are often given for one megohm circuit impedance, an amplifier band-width of 200,000 cycles, and in peak-to-peak maximums. The possibility of using a chopper at that impedance, with a following very wide-band amplifier, which will respond to peak values, gets rather remote.

Figure 31, illustrates the approximate relationship existing in 3 popular type numbers.

We put our amplifier into operation and with no great surprise, we observe there is output with zero signal input. Where should we look for the noise?

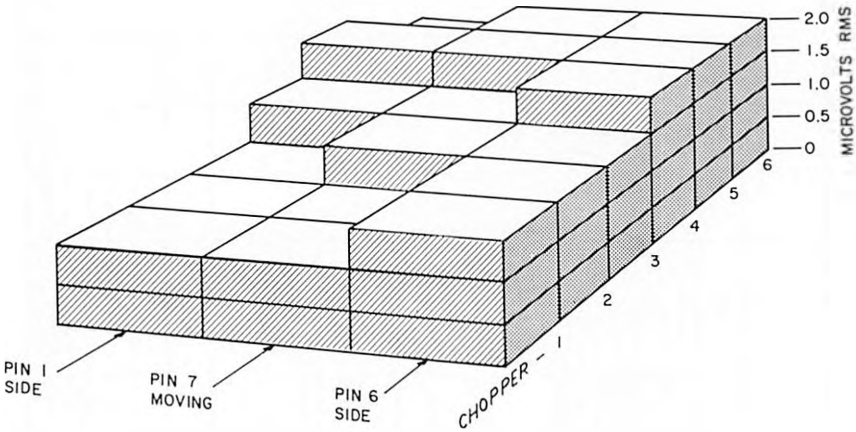


FIGURE 29. *Measured noise levels of six Airpax type 2300 choppers.*

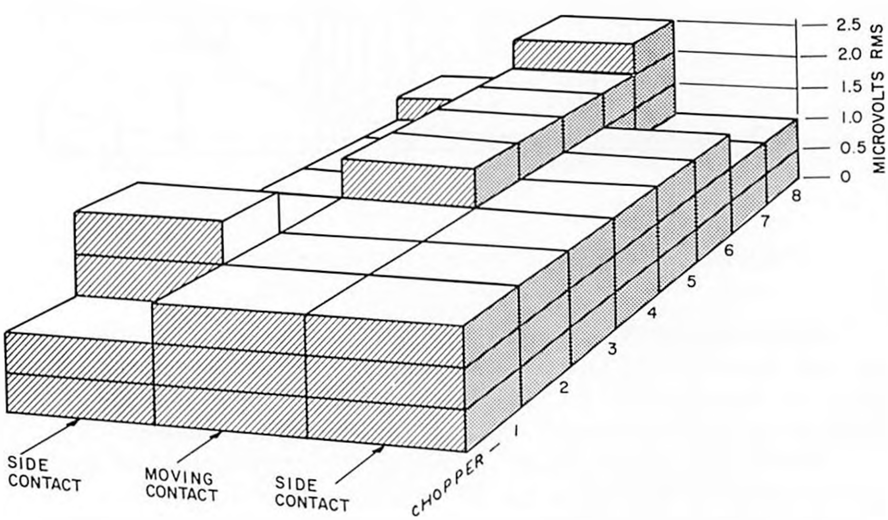


FIGURE 30. *Measured noise levels of eight choppers, manufacturer B.*

The answer depends on the amplifier, so let's assume a transformer and chopper input with a chopper demodulator and an overall gain of 1000. We find that 250 millivolts of offset appears. The chopper has a 400 cycle drive, and the amplifier, a band-width of about 100-2000 cycles and an input impedance of 10,000 ohms. The AC noise wave form is examined, ahead of the demodulator, and appears to have a definite pulse shape. The points of input chopper make and break are clearly visible on the oscilloscope.

Let's pull the input chopper out of the socket. The noise reduces

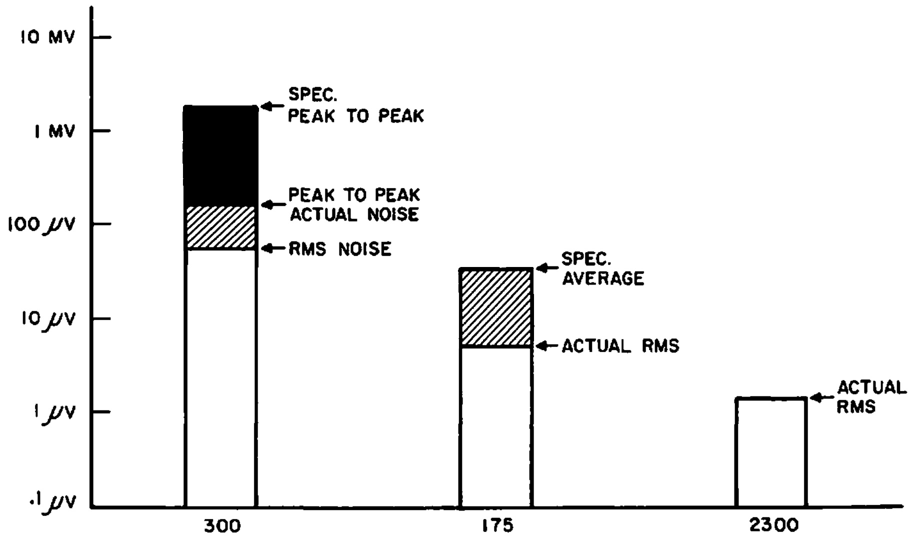


FIGURE 31. *Chopper noise limits are sometimes given as the worst possible combination.*

sharply, although some DC output remains. The pulse shape vanishes, leaving some 400 cycle component, badly distorted.

At this point we can deduce:

1. The equivalent input is 250 microvolts, amplifier bandwidth is restricted, and circuit impedance only 10,000 ohms, so at the worst only a small fraction of the noise can be the fault of the chopper.
2. The definite pulse shape of quite high level indicates DC leakage near the input. It might be a thermal junction, but at 250 microvolts this isn't too likely.
3. There is also hum induced in the input stage or stages, possibly in the input transformer.

Now, noise is a package with many wrappings. As we peel off one or two layers we get the box smaller, but our problem becomes more complex and the right answer more subtle. Some of the possible sources of noise are

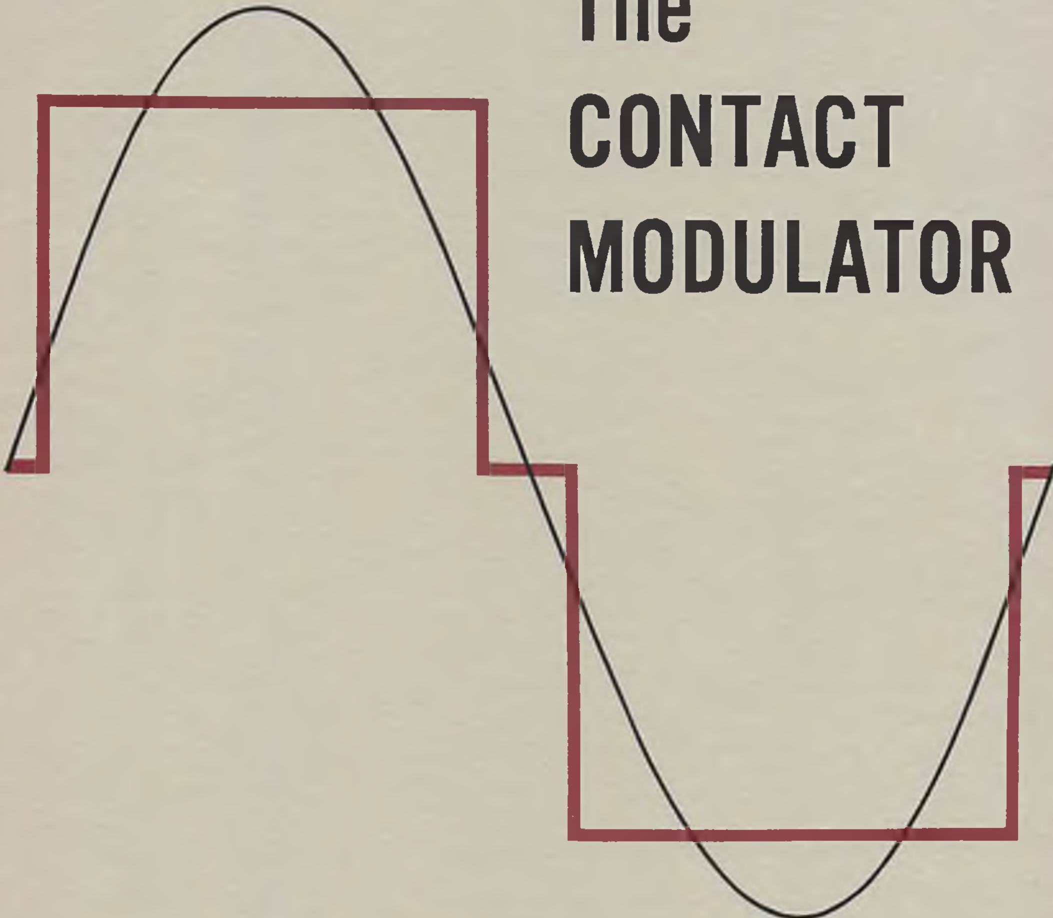
tabulated in the following listing. (Did you try the amplifier without the chopper)?

- DC leakage from the power supply to the chopper input.
- AC leakage, resistive, or capacitive, or inductive coupling to the input circuits, from power wiring or filament lines.
- DC fluctuation in the amplifier power supply. (Even a very slow change may produce a potential which will be chopped).
- Insufficient hum filtering in the DC supply.
- Hum pickup in the input transformer or wiring from the power supply chokes and transformers.
- Microphonic input tubes.
- Static noise, or microphonic noise, from capacitive changes in the wiring due to vibration and shock.
- (At high impedance), slowly changing grid current, giving a signal to be chopped.
- Poor quality (leaky) coupling capacitor at the chopper input.
- Thermally produced potentials at joints, from solder or dissimilar metals.
- Contact noise in resistors.
- Hum pickup in wound resistors.
- Common ground loops.
- Hum loops induced in the chassis.
- Random or Johnson noise (which can have any frequency other than zero).
- Hum or noise from the input tube or transistor.
- And, sometimes, a noisy chopper.

Finally, with the above and other variables, do not forget that any change, seeming to be an improvement, may add hum which cancels that already on hand. Such temporary improvement almost certainly leads to trouble if more than one amplifier is involved.

AIRPAX

The CONTACT MODULATOR



PART 6: Chopper Amplifier Design

THE SIXTH OF A SERIES

AIRPAX

THE CONTACT MODULATOR

PART VI

CHOPPER AMPLIFIER DESIGN

WRITTEN BY

THE ENGINEERING STAFF

OF

AIRPAX ELECTRONICS

INCORPORATED

Cambridge, Maryland • Fort Lauderdale, Florida

H. A. COOK, President

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TABLE OF CONTENTS

SECTION I

FREQUENCY RESPONSE IN CHOPPER AMPLIFIERS

Choppers have a number of rather sophisticated uses, and later we will detail some of these circuits, such as their application as filters, in noise generators, in the control of primary standards of frequency or in VHF direction finding equipment. However, the first (and still most prevalent) use, has been as the modulating element in DC amplifiers.

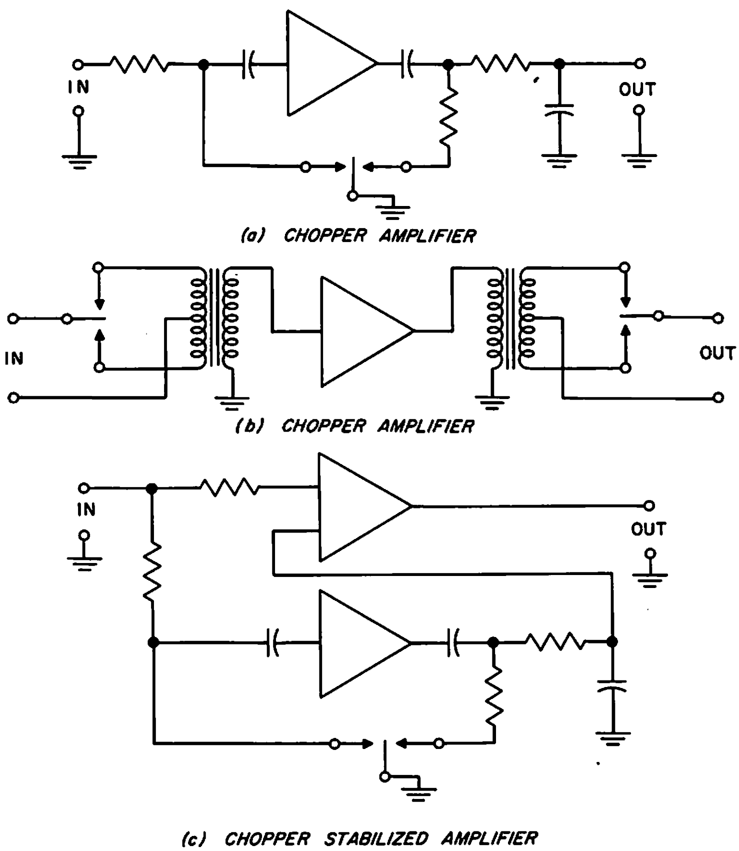

FIGURE 1. *DC amplifiers using choppers are of 3 basic types.*

The output of many transducers is DC, and if not an absolute requirement, it will usually be found easier to use DC than to design around the obstacles presented by AC powering the transducer. Also, most mathematical functions are more easily performed with DC. In even the simple

operation of summing two signals, it is obviously easier to add DC values than to insure the phase relationship of two AC signals, even powered from the same mains.

Three basically different types of modulated amplifiers have evolved. The simplest is commonly called a chopper amplifier, [1,2] as in Figure 1a, employing half wave modulation, and providing for demodulation in the same chopper. As Dr. Hutcheon points out in Part III, there is a variety of modulation methods. Figure 1b, is another chopper amplifier, for want of a better term it is described as a chopper-transformer amplifier.[3] It has the unique virtue of providing isolation as well as a differential input, for this reason it is commonly used as a data amplifier, of which much more later. A third variation is the chopper stabilized amplifier,[4,5,6] Figure 1c, employing a chopper amplifier to stabilize the zero and gain of a wide

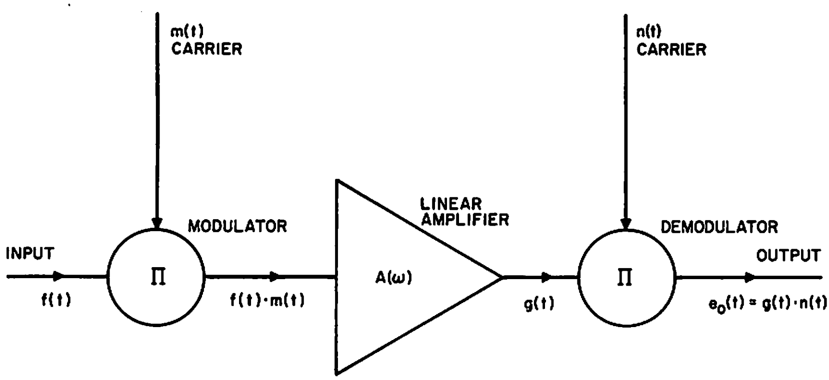

FIGURE 2. *Descriptive terminology used in analysis.*

frequency range amplifier. This last and widely used device is sometimes called the Goldberg [4] amplifier after the man generally credited with its invention. It finds use as an operational amplifier in many computers.

The Goldberg circuit neatly sidesteps one limitation of chopper amplifiers. Obviously, if the response time of the desired signal approaches the vicinity of its carrier, difficulty will be experienced. If the DC component is to be re-established, and if ripple from the carrier (chopper) is objectionable, it may become necessary to filter the demodulated output. In so doing, we often so severely limit the amplifier response time that the carrier frequency is not involved.

[1]Amplifier, J. W. Milnor, U.S. Patent No. 1,378,712, filed January 17, 1918.

[2]DC Amplifier Stabilized for Zero and Gain, Williams, Jr., Tarpley and Clark, AIEE Transactions, p. 47, 1948, vol. 67.

[3]A Contact Modulated Amplifier to Replace Sensitive Suspension Galvanometers, M. Liston, et al, Review of Scientific Instruments, Vol. 17, No. 5, May, 1946.

Even with an output filter, the carrier frequency becomes a limiting factor in data amplifiers, as later discussed. Since the carrier frequency is some sort of limitation, let's explore the subject. The following is a review which will point out certain difficulties of definition and will clear up some popular misconceptions.

In Figure 2, we set out terminology of the basic system. Modulation of the input signal $f(t)$ is accomplished by multiplication with the carrier. This carrier is a periodic function which we designate as $m(t)$. The product is passed through a system which is assumed linear and characterized by its frequency behavior $A(\omega)$, or in more general terms of complex frequency, $A(s)$. The output of $A(\omega)$ we shall call $g(t)$. This is

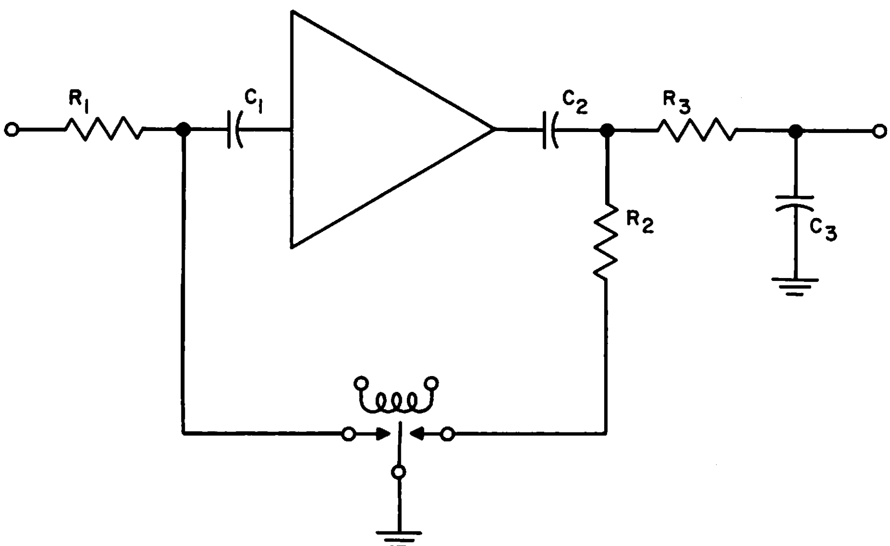

FIGURE 3. *Half wave chopper modulator and demodulator.*

demodulated by multiplying by an output demodulator waveform $n(t)$ to produce the final output $e_o(t)$. We shall confine our attention to the most practical case where one desires $e_o(t)$ to be a faithful but amplified reproduction of $f(t)$.

Very often $m(t) \equiv n(t)$. One conspicuous exception is the half wave chopper modulator circuit of Figure 3, in which $m(t)$ and $n(t)$ are complimentary functions. Unless specifically excepted, we shall assume $m(t) \equiv n(t)$, both having a basic frequency of ω_0 radians per second.

[4]Stabilization of Wide-Band Direct Current Amplifiers for Zero and Gain, Goldberg, RCA Review, p. 296, June, 1950.

[5]Driftless DC Amplifiers, Bradley & McCoy, Electronics, April, 1952.

[6]Stabilized Direct Current Amplifier, E. A. Goldberg, Assigned to RCA, U.S. Patent No. 2,684,999, filed April 28, 1949.

Our first example will be a trivial one in which we shall achieve perfect fidelity, that is, $e_o(t) \approx f(t)$. To do this, let $m(t) \equiv n(t)$ be a perfect square wave as shown in Figure 4a. Let $A(s) \equiv k$, a real constant. Then:

$$e_o(t) = f(t)\, m(t) k\, m(t) = k\, f(t) \quad \textit{since} \quad m(t)\, m(t) = m^2(t) \equiv 1$$

It should be remarked that this is a perfect carrier amplifier. The result is trivial since $A(s) \equiv k$ implies a perfect DC amplifier in which

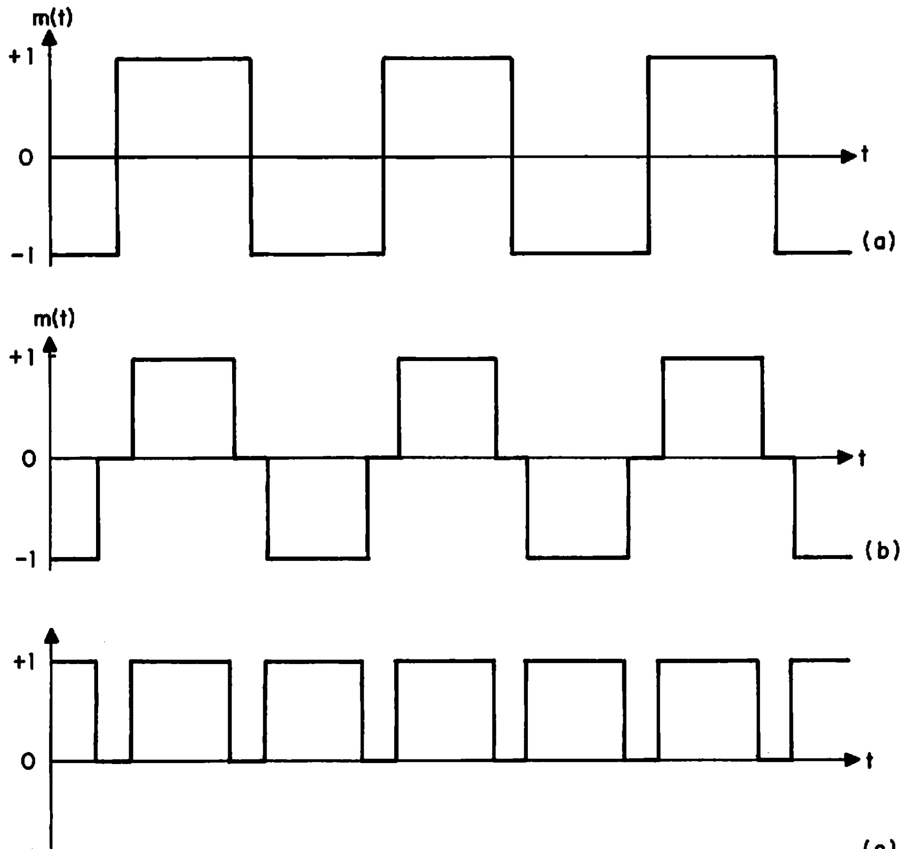

FIGURE 4. *Chopper system waveforms.*

case the whole process of modulation and demodulation becomes pointless. Nevertheless, by example, it disproves the popular notion that no carrier amplifier can pass more than half its sampling frequency.

It is worthwhile here, for comparison later, to investigate this system in the frequency domain rather than the time domain. The procedure

might be as follows:

$$f(t) = \frac{1}{2\pi}\int_{-\infty}^{+\infty} F(\omega)\, e^{\,i\omega t}\, d\omega \qquad\qquad m(t) = \sum_{-\infty}^{+\infty} c_n e^{\,in\omega_o t}$$

where $F(\omega)$ is the Fourier transform of $f(t)$ and the c_n represent the complex coefficients in the Fourier expansion of $m(t)$.
Then:

$$f(t)\, m(t) = \frac{1}{2\pi}\sum_{-\infty}^{+\infty}\int_{-\infty}^{+\infty} c_n F(\omega)\; e^{\,i(\omega + n\omega_o)t}\, d\omega$$

To picture this along the frequency axis, we show in Figure 5, the spectral power densities of some hypothetical input $F(\omega)$, the square wave $m(t)$ or $M(\omega)$ and the spectrum of their convolution corresponding

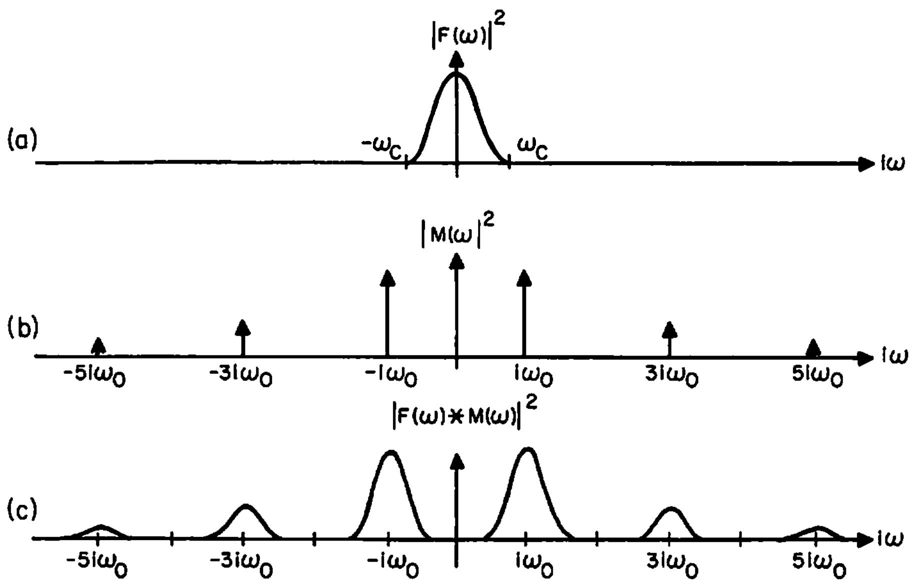

FIGURE 5. *Frequency components arising from modulation.*

to multiplication in the time domain. In words, all the frequency components in $f(t)$ beat against all the frequency components in $m(t)$ creating sidebands. If $f(t)$ is band limited so as to constitute an island spectrum about DC, then beating against ω_o and all the harmonics (odd only in this case) in $m(t)$ will recreate these island spectra about each harmonic of $m(t)$ weighted by the appropriate coefficient c_n.

If ω_c is the highest frequency component in the band limited signal $f(t)$, it can be seen that if $\omega_c > \omega_o$, adjacent islands in the product spectrum

will start to overlap. At this point, one is tempted to say that frequency components lying in overlap regions are hopelessly scrambled in the sense that one cannot tell which island spectra they came from so that they cannot be reassembled by detection. This would represent a loss of information leading to the conclusion that $f(t)$ must be band limited to the modulator frequency. We note that the modulator samples the signal $f(t)$ twice per cycle so we arrive at the maxim of information theory that $f(t)$ must be band limited to one-half the sampling rate.

From our conclusions in the time plane, this statement is incorrect. The reason is that frequency components in overlap regions are very definitely phase correlated and demodulation (which is phase detection) can unscramble these overlap regions with no loss in information. As for the sampling rate theorem, it is inapplicable since the square wave of $m(t)$ is doing much more than simply sampling at discreet points in time.

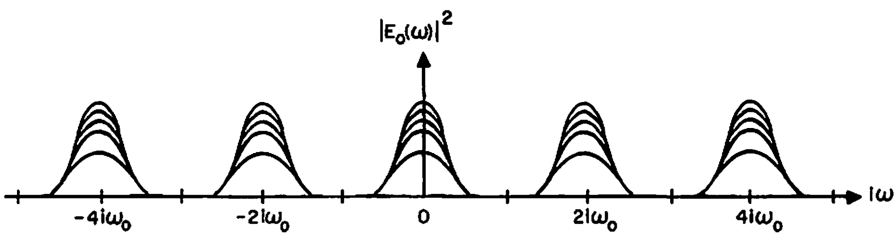

FIGURE 6. *Spectrum of detected output.*

Proceeding with demodulation, we multiply again by $m(t)$. All the island spectra of Figure 5c are beaten down by all the harmonics of $m(t)$ in Figure 5b to form Figure 6. This is composed of island spectra at all the even harmonics of ω_o, each island is the limit of an infinite series of islands.

Analytically,

$$e_o(t) = K\ f(t)\ m(t)\ m(t) = \frac{K}{2\pi} \sum_{j=-\infty}^{+\infty} \sum_{n=-\infty}^{+\infty} \int_{-\infty}^{+\infty} F(\omega) c_{-n}\ c_{n+j}\ e^{i(\omega + j\omega_o)t}\ d\omega$$

This uncomfortable equation is actually a false front.

Consider:

$$m(t)\ m(t) = \sum_{-\infty}^{+\infty} c_n e^{in\omega_o t} \sum_{-\infty}^{+\infty} c_m e^{im\omega_o t} = \sum_{\substack{j=-\infty \\ n=-\infty}}^{+\infty} c_{n+j}\ c_{-n}\ e^{ij\omega_o t}$$

which is a doubly infinite summation similar to that shown in Figure 6. Yet we know that $m^2(t) \equiv 1$. Evidently then:

$$\sum_{n=-\infty}^{+\infty} c_n\ c_{-n} = 1 \qquad\qquad \sum_{n=-\infty}^{+\infty} c_{n+j}\ c_{-n} = 0 \quad \textit{for all } j \neq 0$$

and this is actually quite true if one goes to the trouble of evaluating the

various c_n for a square wave. This means that all the island spectra in Figure 6, that are beaten down to DC, add up to $F(\omega)$ while _all_ the other island groups completely cancel each other out. Figure 6, being a power spectrum does not show this cancellation, but in reality Figure 6 is simply of the form of Figure 5a. This should make it clear that the frequency domain representation of the modulation-demodulation process is an extremely awkward one compared to the time domain representation. As we shall see, it is much easier to discuss the step response of practical chopper amplifiers than it is to discuss or even define their frequency response.

Departing from this trivial but indicative example, our first step towards realism is to release the requirement that $A(\omega)$ should pass DC since $A(\omega)$, in practice, is an RC or a transformer coupled vacuum tube (or transistor) amplifier which has no inherent null offset. Let us assume then that $A(\omega)$ is flat at high frequencies (at least such that it can pass a step with negligible rise time) and passes low frequencies down to say 10 CPS. Let the carrier frequency be 400 CPS to use a familiar figure. Now if $f(t)$ is band limited to 390 cycles, our modulated wave $f(t)\,m(t)$ contains no spectrum elements in the band 0 to 10 CPS and so is passed completely by the amplifier. In this case $e_0(t) = f(t)$ with no distortion. If $f(t)$ contains spectrum elements in the range 390 to 410 cycles, these will be beaten down within the range 0 to 10 CPS and not passed by the amplifier.

The reassembled signal $e_0(t)$ will consequently be lacking in components near 400 CPS. They will not be absent altogether since there will be contributions beaten down from 800 CPS. All components in $f(t)$ between 410 and 1,190 cycles will be passed without distortion so it is a mistake to say that 390 CPS represents the upper frequency response limit of the system. Between 1,190 and 1,210 cycles we have another notch that is beaten down to DC against the third harmonic of the carrier and not passed by $A(\omega)$. Other notches occur at $[400\ (2n+1)] \pm 10$ CPS. The question as to what the useful bandwidth of this system is should be objected to on the grounds that it is an inadmissible question until one defines bandwidth. Clearly we cannot use any standard definitions such as the -3DB criterion.

A more meaningful subject to investigate is the step response of this system. In the time plane this is a simple question as shown in Figure 7. The leading edge of $f(t)$ is passed _at once_ by the amplifier $A(\omega)$ with the same alacrity with which it passes the square pulses of $m(t)$, were $f(t)$ simply DC. Of course, the pole zeros pairs of $A(s)$ at the origin, will result in some phase shift at 400 cycles and will appear as a slight droop on the output wave form $g(t)$, but if the rejection band of $A(\omega)$ is only 10 CPS this droop will be very small.

This droop is shown exaggerated in $e_0(t)$ of Figure 7. If one attempts

to determine the nature of $e_0(t)$ on a spectrum basis, the convolutions in the s plane, complicated by the weighting of $A(s)$, will soon lead to an integral nightmare. Avoiding this, for all practical purposes, the step response of this system is perfect, being limited only by the rise time of $A(\omega)$ which we have taken as negligible. The conclusions to be drawn are that one should not attempt to specify the "frequency response" of such a system and that carrier amplifiers need not be limited in their response time by any consideration of the carrier frequency.

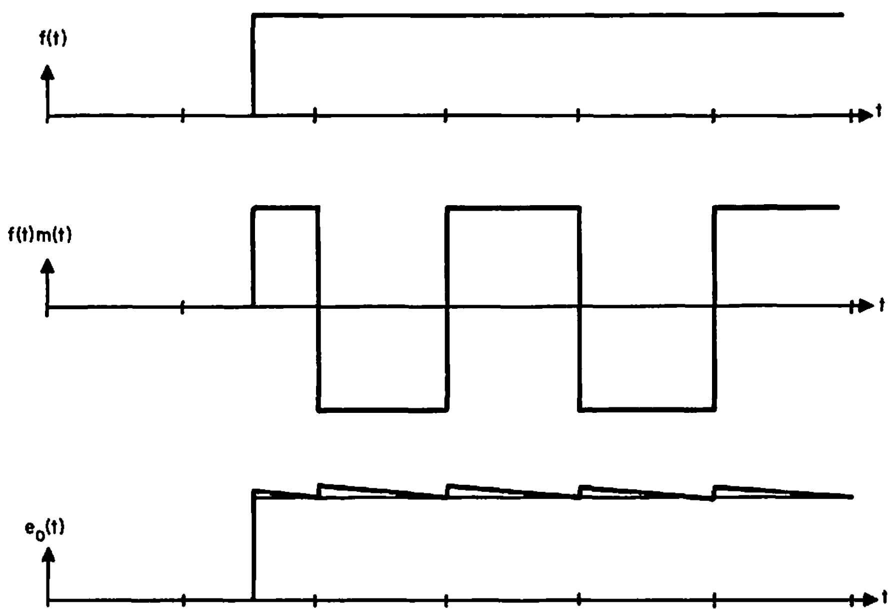

FIGURE 7. *Step function with resultant performance.*

Finally, we should relieve the last ideal assumption, that of specifying $m(t)$ as a square wave. In chopper amplifiers, at least, the function $m(t)$ will appear as shown in Figure 4b. We shall still assume that $m(t) \equiv n(t)$ through adequate phasing. If we revert temporarily to the assumption that $A(\omega) \equiv 1$ then:

$$e_0(t) = f(t)m^2(t)$$

where $m^2(t)$ is shown in Figure 4c. Consequently $e_0(t)$ is a reproduction of $f(t)$ with 800 CPS rectangular notes punched in the wave form. These windows are, of course, most objectionable if one requires fidelity. Music played through such an amplifier is not a little obscured by the 800 cycle howl that these holes produce. If we re-establish the zero at DC in $A(\omega)$

and inquire as to the frequency response, the 800 cycle component and its harmonics appear as a failure of all the island spectra being beaten down to 800 CPS by demodulation to cancel each other out.

The step response most clearly illustrates the loss in information due to the off time. If the step occurs during the on time, it is passed immediately by the entire system (assuming good high frequency behavior of $A(\omega)$). If the step occurs in the off time, it is delayed by, at most, the time

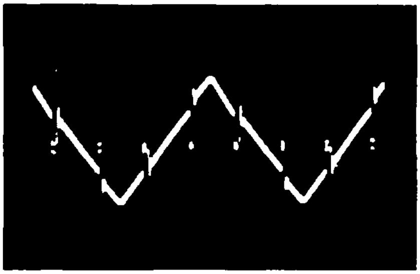

FIGURE 8.

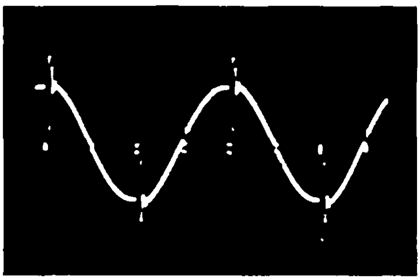

FIGURE 9.

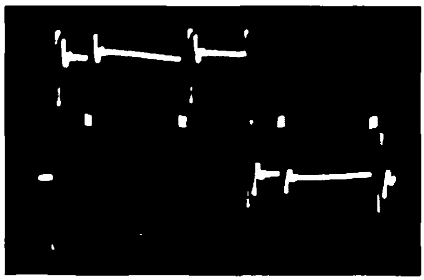

FIGURE 10.

Oscillograms of $e_0(t)$ with $f(t)$ a triangular wave (Figure 8), sine wave (Figure 9), and square wave (Figure 10).

duration of the off time. Thus in a 400 CPS chopper amplifier whose off time is 10% of its on time, the step response is at most .125 milli-seconds.

In Figures 8, 9 and 10, we have oscillograms of $e_0(t)$ when $f(t)$ is respectively a triangular wave, sine wave and square wave, all at a frequency of 200 CPS. The latter figure illustrates the step function response of the system since a square wave is decomposable into a linear combination of delayed step functions. The pass band of the amplifier $A(\omega)$ was 2 CPS to 180 KC. Figures 11 and 12, respectively show the effects of decreasing the upper band limit to 1.5 KC and increasing the low pass limit to 80 CPS when $f(t)$ is a 200 cycle square wave.

Figure 11, illustrates the point that the step response of the system is more dependent on the bandwidth of $A(\omega)$ than it is on the chopper frequency. Figures 13 and 14, show the difficulties encountered when $f(t)$ is a 400 CPS sine wave. The two pictures simply show encounters at two different phase relationships between $m(t)$ and $f(t)$. It is clear that a 400 CPS square wave has been removed from $e_o(t)$ creating harmonics and a loss of fundamental. This is due to the failure of $A(\omega)$ to pass DC. Figure 15, shows that the system passes an 800 cycle sine wave with no distortion other than the 800 CPS windows punched in the wave form. Figure 16, shows the trouble when $f(t)$ is a 1.2 KC sine wave. Some of this signal is beaten down to DC by the third harmonic of $m(t)$ and cannot be passed by $A(\omega)$. Again, the result is to extract a 400 CPS square wave from the output.

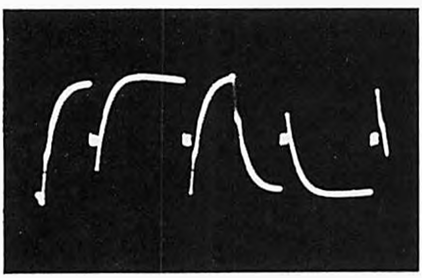

FIGURE 11. *Pass band 2 CPS to 1.5 KC.*

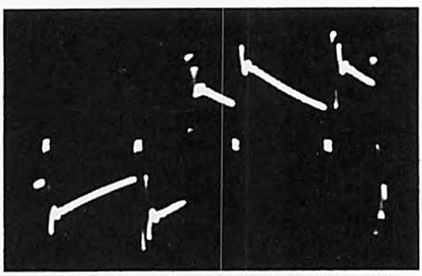

FIGURE 12. *Pass band 80 CPS to 180 KC.*

One cannot ignore what is to be done with $e_o(t)$ for overall frequency response. Since $e_0(t)$ only has meaning as an average value (averaged over the 800 CPS holes punched in it), it is inevitably filtered whether the filter is a meter needle, pen recorder, or a more complicatd linear network. One is filtering chiefly against the 800 cycle component and sidebands coming down spectrum from it. If $f(t)$ is DC, the output is as shown in Figure 4c. One can view the 800 cycle windows as 800 cycle spikes upside down. The weight of the 800 cycle component in the Fourier spectrum of these spikes is $(2/\pi)\sin 2\pi\,(T/P)$, where T is the off time width and P is the period of an 800 cycle train. Thus, if the off time is 10% of 1.25 milliseconds, the ratio of unwanted 800 cycles to the DC signal is $(2/\pi)\sin 36^\circ = .374$.

Notice as $T \rightarrow 0$, the 800 cycle component vanishes as in the second system we considered. Now .374 is —8.6 DB. If —20 DB is acceptable with which to reject the 800 cycle ripple, we need —11.4 DB more. For a single time constant filter rolling off at —20 DB/dec. we may spot its —3DB cutoff frequency at 229 cycles which is now the overall bandwidth. For a double time constant filter (say a Butterworth double), the

overall bandwidth could be extended to 420 cycles. Consider a 400 cycle chopper with only 5° of off time. Then $(2/\pi) \sin 10^\circ = .11 = -19.2$ DB. In this case, one could do away with filtering altogether, that is, use a transducer whose rise time is of the order of the off time which is now 34.7 microseconds, corresponding roughly to a pass band of 4.6 KC.

We thus see that the response of a chopper amplifier is much more intimately related to the off time than to the carrier frequency. Such a system will not, of course, amplify sine waves of frequencies which are odd multiples of the carrier frequency, but one seldom cares to do this in any event. Rather one wishes to amplify more complex signals having only small power densities located in the immediate region of these odd

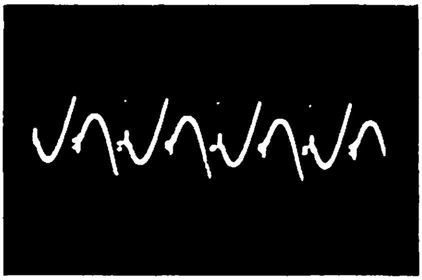

FIGURE 13. *400 CPS input.*

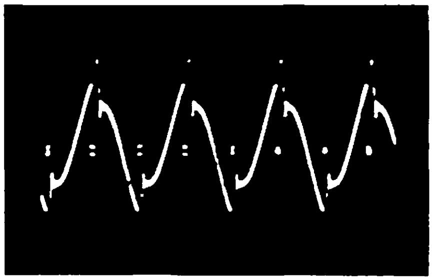

FIGURE 14. *400 CPS input.*

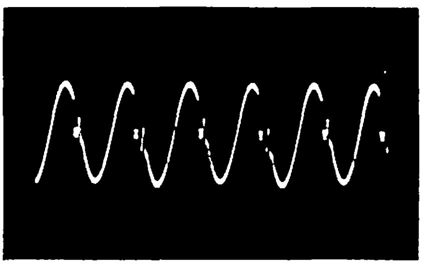

FIGURE 15. *800 CPS input.*

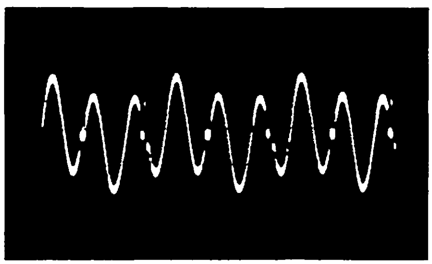

FIGURE 16. *1200 CPS input.*

multiples. It also becomes clear that the amplifier $A(\omega)$ should have a good low frequency response to keep these spectrum notch widths at a minimum.

In summary then, we have the following considerations. The AC carrier amplifier should be designed to have a wide bandwidth, negligible droop and a fast rise time. The chopper off time, representing pure information loss is much more important in bandwidth considerations than is the carrier frequency. It is much more meaningful to discuss the rise time of a chopper amplifier than the frequency response which is difficult to even define usefully.

SECTION II

CHOPPER AMPLIFIER CIRCUITRY

Mechanical choppers are frequently used as demodulators for the identical reason they find use as the modulator. Considered as a multiplying device, they are uniquely capable of approximating periodic multiplication by factors of 0 and 1, at the same time they do not introduce offset. No other device so closely realizes this ideal,[7] although many approaches can be made.[8,9] The usual reason is a natural engineering desire for the elimination of mechanical devices, although choppers have a probable life much in excess of 5000 hours.

A frequently used method employs half wave input modulation, with a SPDT chopper as shown in Figure 1a, thus permitting demodulation with the unused contact. Usually the moving contact is at or near ground potential. Since input and output are necessarily brought together, some capacity between input and output necessarily exists, and spurious oscillation is possible with high gain amplifiers. The remedies are several:

1. The high frequency response of the amplifier can often be attenuated. Since the interelement capacity of a chopper will be low (about 1 millipicofarad for an Airpax Type 300), and the stray coupling capacity external to the chopper can be held to a minimum, the upper pass band can often be limited without greatly disturbing the rise time of the square wave carrier. It is obviously preferable to ground the moving arm to provide shielding between fixed contacts, and choppers are internally arranged to promote this shielding.

2. An odd number of amplifier stages can be used, making any feedback present negative instead of positive. Note that this will also reverse the sign of the output DC, a characteristic which might be either good or bad, depending on the ultimate use.

3. Demodulation can be accomplished in another chopper (or by other methods). At first glance this immediately raises the question of "tracking", i.e., will the two choppers switch synchronously? In practice, as mentioned in other Parts, the electrical linkage is stiffer than a mechanical link. Two choppers, each SPDT, will run as close or closer than one DPDT unit, and is obviously more flexible. Naturally, the chopper design should be the same, and for maximum accuracy the two should be purchased as one unit, Figure 17, or specified in pairs.

[7]D-C to A-C Modulators, George Sideris, Electronics, p. 47, January 23, 1959.

[8]Switching Transistors Used As A Substitute for Mechanical Low-Level Choppers, Communications and Electronics, pp. 55-157, March, 1955.

[9]Servo Modulators, Barber & Klivans, Control Engineering, Oct.-Dec., 1957.

The Philbrick* Type K2P amplifier has had considerable popularity for some years, Figure 18. It uses a 12AX7 twin triode AC amplifier, and is modulated and demodulated by an Airpax Type 175 chopper operating at 60 cycles. A long time constant output filter holds the response time to a fraction of a cycle per second. The unit is sometimes employed as an operational amplifier directly, or as the stabilizing unit of a wide range DC amplifier.

A similar unit is sold by Electrol** as their Type 1C, the circuit for which is shown in Figure 19. The input impedance is about 2 megohms. The output is filtered, establishing the amplifier response time in terms of the RC filter employing a 22 megohm resistor and 1 mfd capacitor.

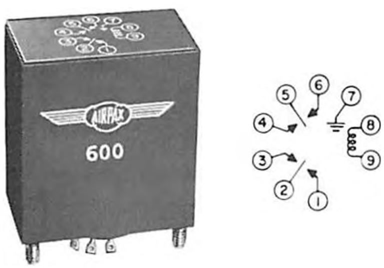

FIGURE 17. *Airpax Type 600 DPDT chopper has complete isolation by use of two choppers coupled electrically.*

FIGURE 18. *Philbrick chopper amplifier type K2P.*

The neon bulb is part of an alarm circuit to signal overload of the associated Type 1C operational amplifier. The overall voltage gain is 1000, with an output of $\pm$ 10 volts DC.

For high stability DC amplification, such as is required for data reduction and control systems, the straight chopper amplifier with transformer input provides certain performance characteristics found in no other DC amplification system. As contrasted with the chopper stabilized amplifier, the straight chopper amplifier converts the input signal to a square wave before amplification. This has some advantages. Since an AC signal is to be amplified, transformer coupling gives electrical isolation of the input circuit from the amplifier. The result is good rejection of common mode signals. The term "common mode," usually refers to a

*George A. Philbrick Researches, Inc., Boston 16, Massachusetts
**Electrol, Inc., 9000 West Pico Boulevard, Los Angeles, California

differential input having an unwanted signal, i.e., hum or noise, applied equally to both input terminals. See Part V.

With no input signal applied to the chopper, there is no signal applied to the amplifier input and no amplifier output in the absence of input, with the exception of noise and hum introduced by the chopper itself or otherwise coupled to the input circuit. The use of another chopper for demodulation re-establishes the original signal without offset. This is also an easy way of obtaining synchronous operation; if another demodulator system is used, it would probably be necessary to introduce phase shift.

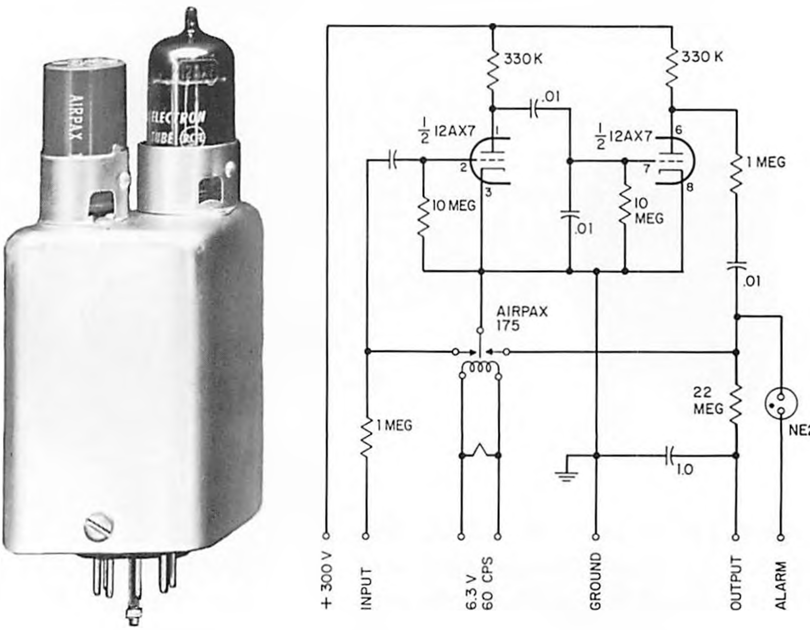

FIGURE 19. *Electrol Type 1C chopper amplifier.*

Chopper amplifiers have been developed by Offner Electronics, Inc.[10] for a number of DC amplification purposes. The accuracy of this amplification system permits it to operate the galvanometers of direct-writing oscillographs. While simple chopper amplifiers were capable of accuracies of about one percent, further refinements permit the high degree of accuracy desired for industrial data logging and control amplifiers. Such accuracy has been obtained in the Offner Type 190 Data Amplifier, shown in Figure 20. (Offner Electronics, Inc., 3900 River Road, Schiller Park, Illinois).

The essential features of the amplifier will be evident on inspecting the circuit, Figure 21. The input and output choppers are mechanically

[10]Recorder Amplifier, Franklin F. Offner, U.S. Pat. No. 2,688,729, filed July 28, 1949.

synchronized, and operate at 400 CPS, from a separate oscillator source. The amplifier is designed to transmit the square wave signal with excellent preservation of wave form. To this end, the amplifier is push-pull throughout, eliminating the need for cathode bypass condensers, which would tend

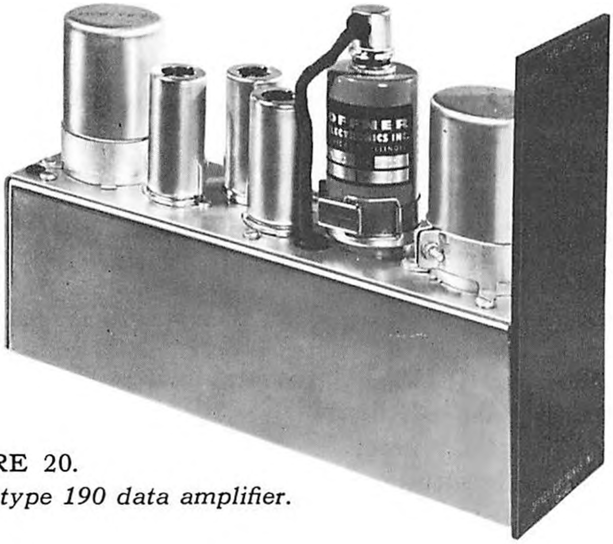

FIGURE 20.
Offner type 190 data amplifier.

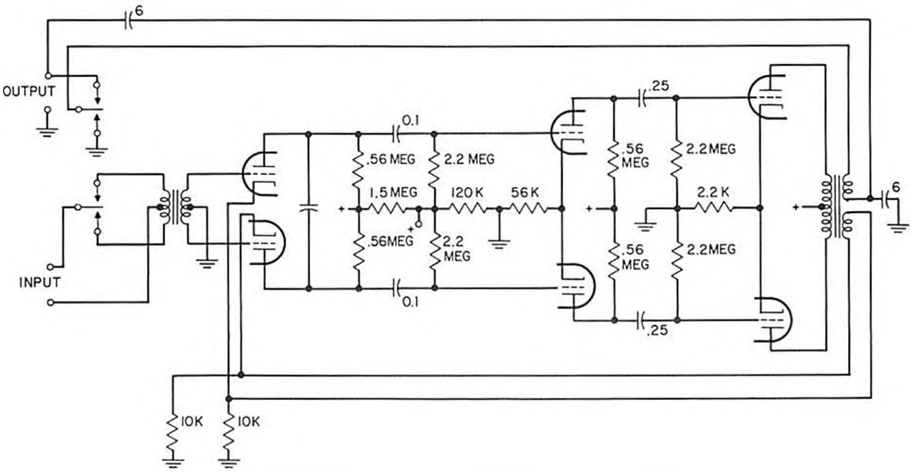

FIGURE 21. *Schematic of Offner type 190 amplifier.*

to give frequency distortion. The output chopper works into a peak charging circuit which produces a DC output signal very precisely equal to the amplitude of the output square wave. Thus the DC amplification is closely equal to the AC amplification factor of the amplifier. The latter

is held constant by employing a large amount of AC feedback, from the tertiary of the output transformer back to the first stage. The use of AC feedback eliminates the need of any electrical feedback from the output circuit, or to the input circuit itself. Thus these may both remain electrically isolated, giving the possibility of true differential input, as well as output (the ground connection shown on the output circuit of Figure 21 may be eliminated, if desired).

It will be seen that no resistors are used in the feedback path. Thus variations in resistor value with time and temperature will not affect the gain, which is primarily set by the transformer ratios. With infinite open loop amplifier gain, the DC gain of the actual amplifier would be:

$$G = \frac{\text{input secondary}}{\text{input primary}} \times \frac{\text{output secondary}}{\text{output tertiary}}$$

In actual practice, the gain is close to this theoretical value, and remains constant both with time and with change in ambient temperature. Experience has shown that the gain will remain constant within a fraction of one percent over a temperature range of $-67°F$ to $+165°F$, and for long periods of operation. Assuring this gain constancy is the fact that the gain is not critically dependent on chopper adjustment. Dwell time and balance may vary several percent without seriously affecting the amplifier performance. In actual practice, this is a most important characteristic. While choppers have proved to be highly reliable when so employed, it must be assumed that their adjustment will always vary a few percent in use over a period of time, and any circuits in which such change causes an appreciable variation in characteristics, cannot be depended upon for long-time stability. Expressed in another way, it is preferable to avoid dependence on the average value of the chopper square wave.

As previously mentioned, the electrical isolation of the amplifier input from the signal input circuit should result in independence of the amplifier output from common mode input signals. In practice, a hundred volts DC can be applied between the input source and ground without producing a measurable output, while a differential signal of one microvolt remains measurable, the rejection of common mode DC signals being therefore better than 100 million. For common mode AC signals, the rejection to AC is not as good, as some small capacitive coupling exists. At 60 cycles a common mode rejection ratio of about one million is practical.

It is of interest to compare the high performance capabilities and relative simplicity of the above chopper amplifier with other available systems of DC amplification. The most straightforward approach is the simple differential DC amplifier, Figure 22. Even with greatest care, however, such an amplifier can hardly be made to come within a factor

of a thousand of either the stability or balance required for data reduction purposes.

The chopper stabilized amplifier, Figure 1c, can be made to have acceptable stability, the drift of the output amplifier being largely degenerated out, and can have rapid response. However, the need to carry DC feedback to the input, makes the amplifier fundamentally single-ended. As such, it is not usable for data logging in many instances. One

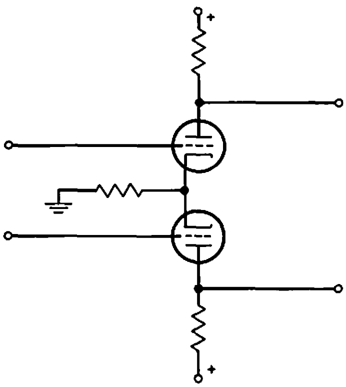

FIGURE 22. *Differential DC amplifier.*

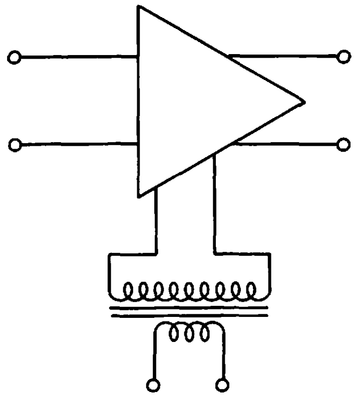

FIGURE 23. *"Floating" the amplifier off ground.*

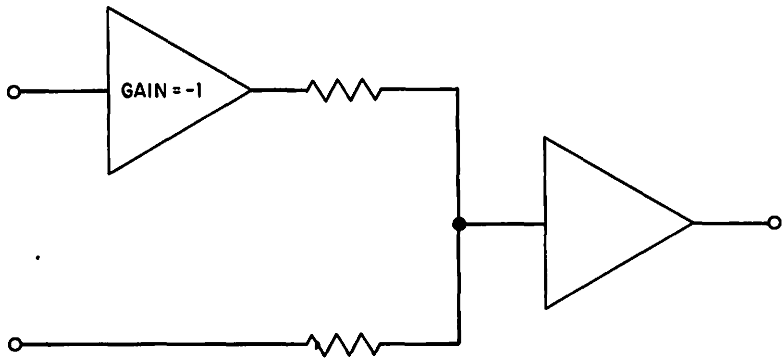

FIGURE 24. *Phase reversing amplifier for differential input.*

approach to making it so usable is to "float" the complete amplifier off ground, employing an isolated, well-shielded power supply, Figure 23. However, while the amplified differential signal exists between the output terminals, neither output terminal may be grounded. Most applications require that the output be grounded, and an auxiliary differential output amplifier is required.

Another approach is to employ a phase reversing amplifier with unity gain in one input line, summing this with the other input, and following this with a chopper stabilized amplifier, Figure 24. The deficiencies of this system reside in the difficulty of maintaining the gains of the two branches exactly equal and opposite. Furthermore, such an amplifier is limited in the amplitude of common mode signal it can tolerate, without overloading the phase reversing amplifier, or causing its gain to vary. Except for these difficulties with respect to balance, the amplifiers of Figure 23 or 24, may have acceptable characteristics.

There are many applications for a differential data amplifier. One common application is for the amplification of thermocouples. These may have an output of from six to thirty microvolts per degree Farenheit. Industrial applications often require an accuracy of one degree or better, sometimes with a total temperature difference in the vicinity of one thousand degrees. Thus the long time stability must be within a few

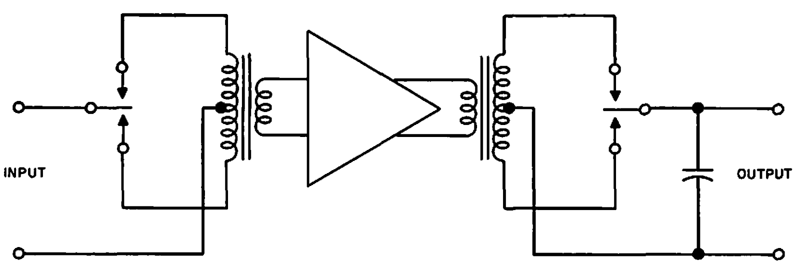

FIGURE 25. *Conventional chopper amplifier with transformer input and chopper demodulator.*

microvolts, with the gain remaining constant to better than 0.1%. To avoid the need for the use of amplifier calibration curves, it is obviously also desirable that the linearity of the amplifier be of this tolerance or better.

The need for differential input results from the possibility that the thermocouple may be connected to some other control or recording device, which may be at a potential difference from ground; or that the nominal ground point of the thermocouple (which may be located hundreds of feet from the amplifier), may have a considerable potential difference from the amplifier ground.

Another common application is for the amplification of strain gage bridge outputs. With an amplifier of this type, DC bridge excitation may be employed, eliminating the need for a reactance balance adjustment. The differential input permits a common DC supply to be employed for a number of bridges simultaneously.

Figure 25, describes the conventional chopper amplifier having a

transformer input. Amplifier response time is improved by short chopper off time and a high chopper frequency, both of which will contribute to ripple reduction and consequent faster amplifier response, by permitting shorter time constants in the output filter, as illustrated in Figure 26.

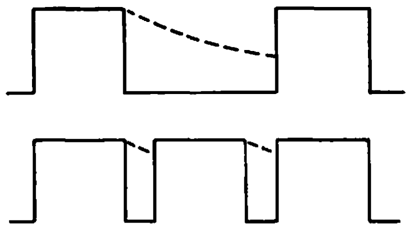

FIGURE 26. *Full wave rectification and short chopper off time improve output filter performance.*

If the amplifier output time constant, composed of the amplifier source impedance and the output capacitor, is made sufficiently low, the amplifier will respond within a half cycle of the chopper excitation frequency. If at the same time, the amplifier load is sufficiently high, the capacitor discharge will be low and the output ripple will be low, permitting fast amplifier response with fairly low output ripple.

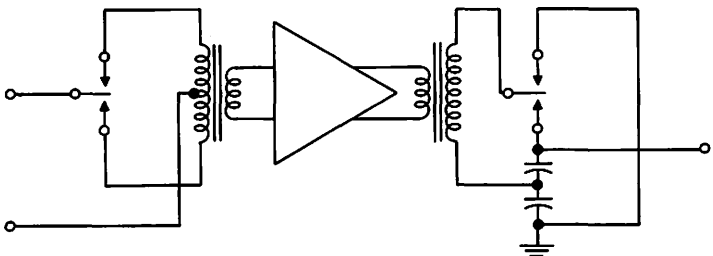

FIGURE 27. *Voltage doubler output used in Offner amplifier.*

The Offner amplifier uses an output voltage doubler, Figure 27, which contributes additionally to ripple reduction caused by unbalance in chopper dwell times, normally specified to be in the vicinity of a 10% maximum difference. Since the voltages across the two capacitors are added separately from each wave half, the output is independent of ripple from that source. Figure 28, shows an exaggerated case of a differing dwell time between wave halves. As the signal passes through the input transformer, a new zero results and on rectification, signals carry ripple contributed by the lack of a perfectly symmetrical dwell.

The use of feedback to improve amplifier performance is customary, and indeed necessary if a high order of gain stability and amplifier linearity is to be obtained. One excellent way of applying feedback around the AC

amplifier circuit is to couple between the output and input transformer, as in Figure 29. If the amplifier responds to peak values of the carrier, then the DC gain is a linear function of the AC gain. Direct DC to DC feedback, as shown in Figure 30, does directly approach the problem, and is most effective in assuring a good degree of stability of gain, and a

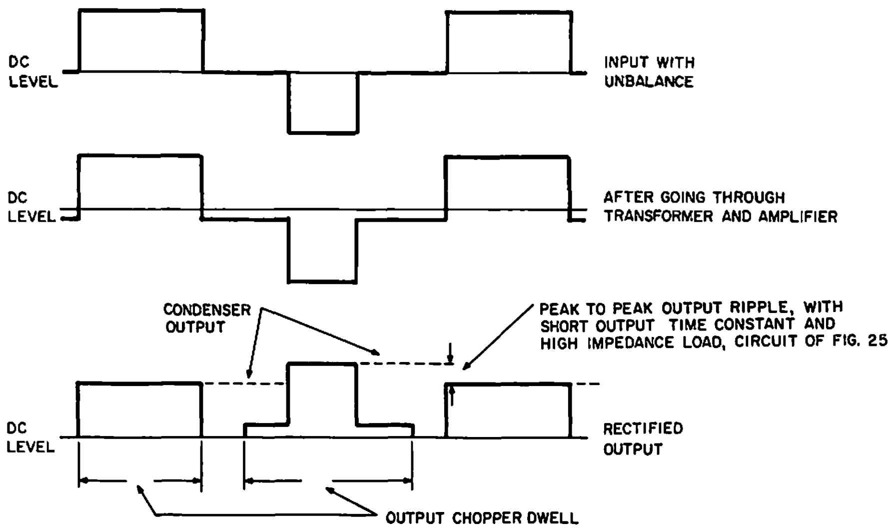

FIGURE 28. *Asymmetrical dwell time.*

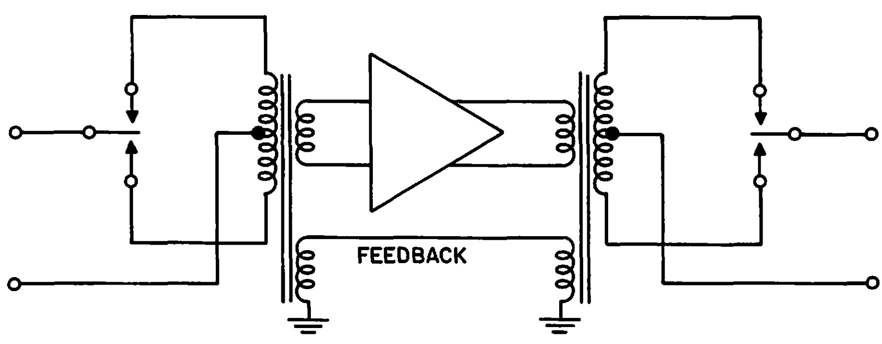

FIGURE 29. *AC feedback in an amplifier.*

high order of amplifier linearity. As will be evident, DC to DC feedback can be supplementary to the AC to AC arrangement of Figure 29.

DC to DC feedback, however, has the unfortunate effect of requiring a common connection between the input and output of the amplifier. The users of strain gages and thermocouples often make life complicated for the amplifier designer by grounding the bridge or couple. Also long

lines become necessary from the transducer to the amplifier, which are subject to stray noise and hum from common mode coupling. A feedback system, ascribed to Dr. Offner,[11] employs two separate and identical feedback windings on the output transformer as shown in Figure 31.

This circuit, in addition to having isolation between input and output,

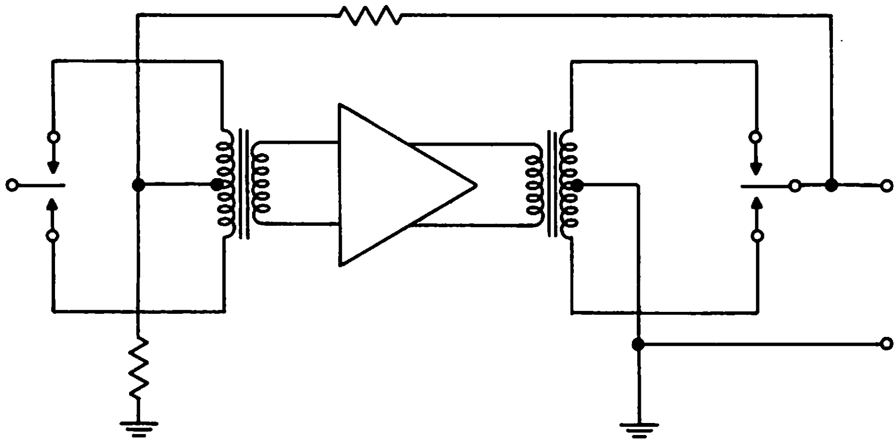

FIGURE 30. *DC feedback across entire amplifier.*

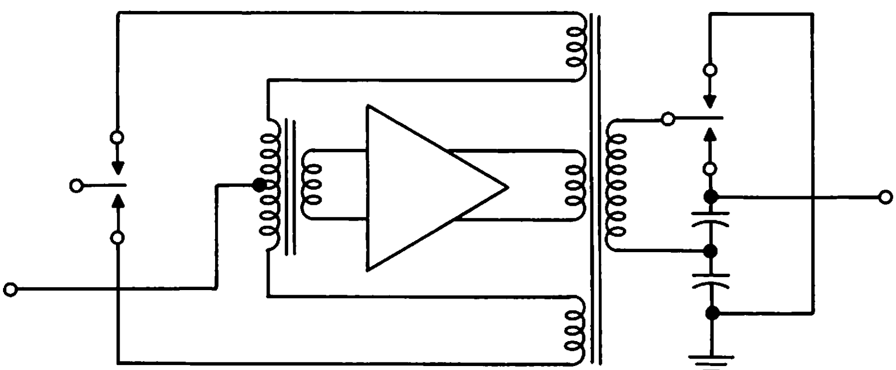

FIGURE 31. *Feedback system combining AC and DC.*

features a high input impedance achieved by the negative feedback arrangement. Although no direct DC feedback is used, a change in input chopper dwell time followed by asymmetry of the "make" points of the output chopper cannot give rise to a change in DC gain as occurs in the circuit of Figure 25. This is based on the assumption of keeping the output chopper dwell time within the input chopper dwell time.

[11] Differential Chopper Amplifier Has High Input Impedance, Dr. Franklin F. Offner, Electrical Design News, March, 1959.

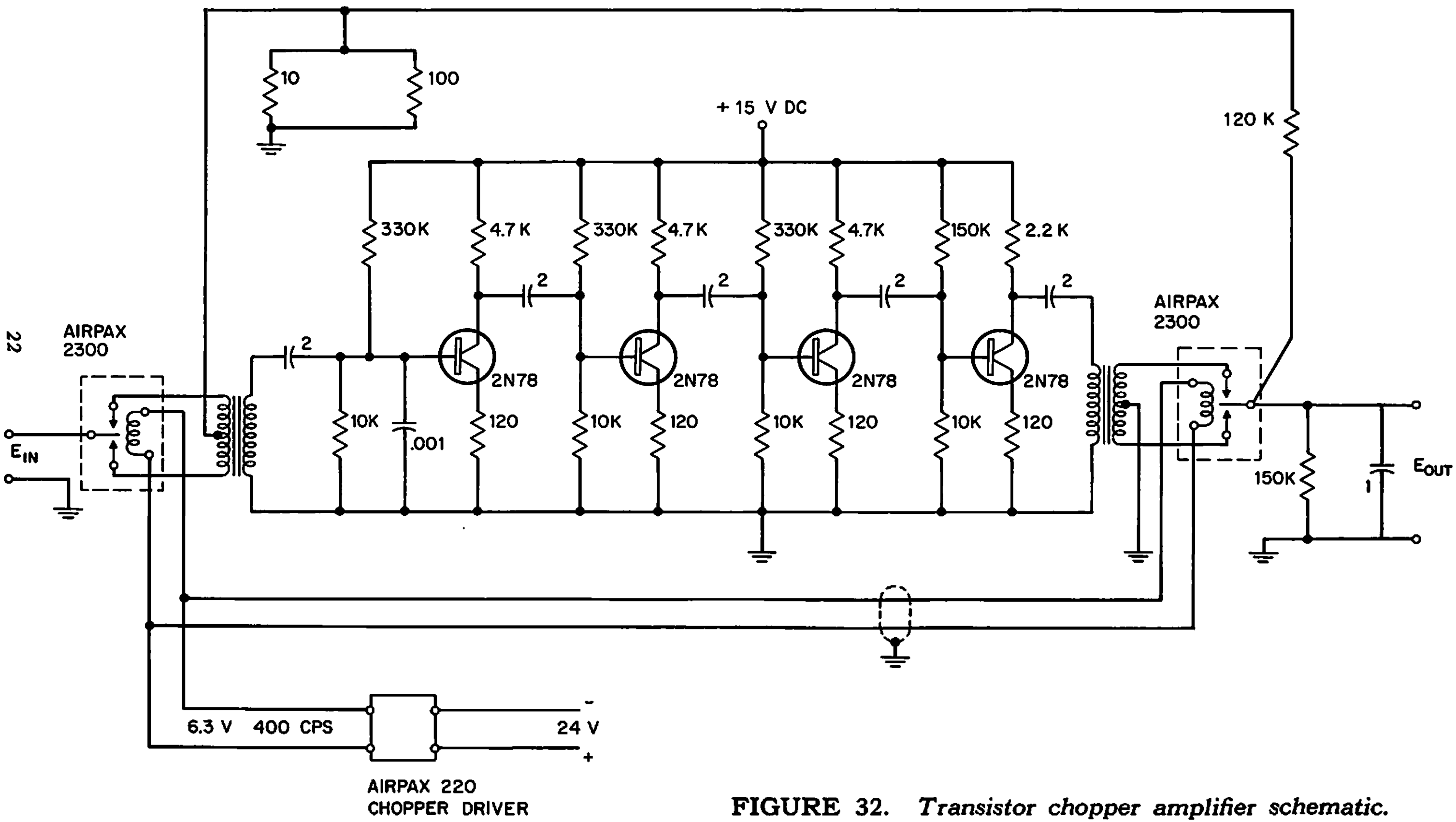

FIGURE 32. *Transistor chopper amplifier schematic.*

The problems associated with the design of a data amplifier are not greatly altered when transistors are used in place of tubes. The input and output transformers, of course, necessarily work into and out of lower impedances than in the case of vacuum tube systems. If high input impedance is required, the stepdown ratio may have to be substantial, resulting in a loss of gain. The transformer size does not shrink merely because one explains to the designer that it will be used with transistors. The input and output transformers may become larger than all the other components. Feedback techniques can be resorted to in an effort to raise the input impedance, but it is difficult to substitute for sufficient inductance in the preservation of low frequency response and the consequent preservation of information.

Figure 32, details the complete schematic of a transistor data amplifier employing choppers. In this particular application, consideration was given to the possible use of transistor choppers. It was desired, however, to measure a signal from thermocouples whose maximum would be about

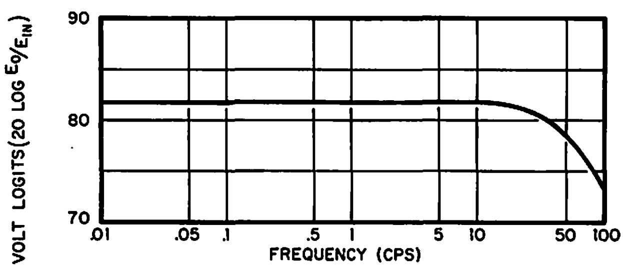

FIGURE 33. *Frequency response of transistor chopper amplifier.*

400 microvolts and to do this with a linearity of 0.1%. It was also desired to recognize a DC signal which might be as low as 4 microvolts. Under these conditions, it proved impossible to use a transistor chopper at the input. The only possible solution appears to be a mechanical chopper. An Airpax Type 2300, a low noise model with leads out the top, was selected for the input. To provide synchronous modulation, another Type 2300 unit was used for the demodulation.

It would have been perfectly possible to use a different chopper at the output, or even to provide some other demodulator. Cases have arisen where the output power requirement was substantial, here we are justified in using a different kind of device or different chopper. This amplifier was expected to deliver ±5 volts DC to telemetry transmitters at a relatively high impedance. Since two identical choppers can be expected to track with a high degree of precision, even under environmental extremes, it was decided to use another low noise Type 2300.

It will be observed that the amplifier frequency response curve, Figure 33, is essentially flat to about 20 cycles using the output filter

described. The filter was only a 1 mfd capacitor, with a 150,000 ohm load. In the application, only DC was available (24 volts DC), and accordingly it was convenient to use 400 cycle choppers by employing an Airpax Type 220 chopper driver. The Type 220 is a transistor inverter designed to drive either 1 or 2 choppers, and delivers 380 to 420 cycles from a DC source of nominally 24 volts. The resultant response is about

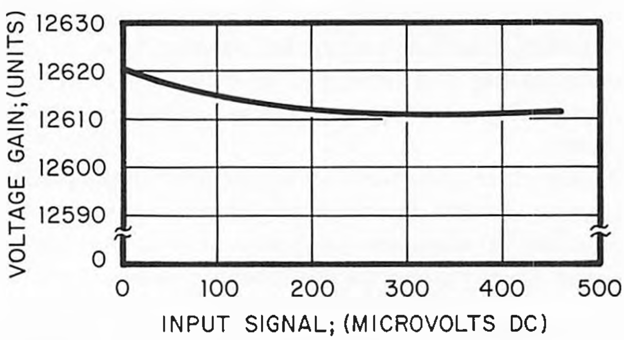

FIGURE 34. *Voltage gain of transistor chopper amplifier.*

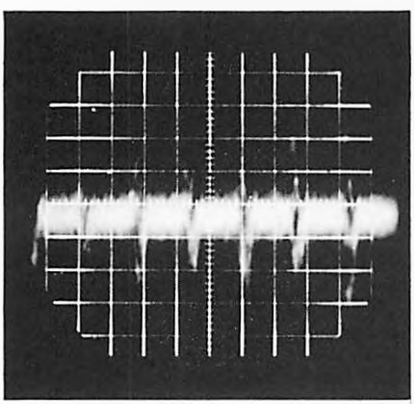

FIGURE 35. *Output of transistor chopper amplifier final stage, with input shorted.*

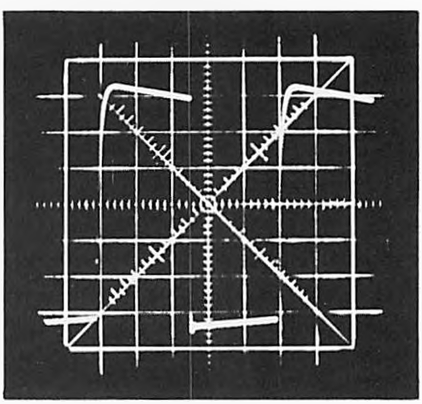

FIGURE 36. *Output of transistor amplifier, 530.4 microvolt input with output chopper removed.*

50 cycles at the —3 DB point, considerably better than the application required, and also considerably poorer than it could have been if a serious attempt had been made to improve response.

The curve of Figure 34, illustrates that a linearity of 0.1% was achieved over most of the dynamic range, and that about a one microvolt offset appears to provide the major deviation from the desired linearity

of better than 0.1%. The ordinate scale is plotted in terms of voltage gain, and 0.1% of a nominal 12610 represents approximately 12623 to 12597. The gain is readily adjustable by changing feedback, as was done in this instance to obtain roughly 12500. No attempt was made to use AC feedback, other than the inherent stage feedback in the transistor circuitry, as it was not necessary for the problem at hand.

The amplifier output observed at the final stage, with input shorted, appears as principally white noise in the oscillogram of Figure 35. The contact action of the modulating chopper is observable, but the 1 microvolt offset is not evident. At such extremely low signal levels (see Part V, Noise in Chopper Circuits), extreme care is necessary to avoid dissimiliar metals which might create spurious thermal electric potentials.

The amplifier delivers ±5 volts to operate telemetry transmitters. The unrectified output appears in Figure 36, displaying a DC input signal of about 400 microvolts. Some droop is observable, representing poor low frequency response. Since the desired performance was reached, no attempt was made toward improvement. The amplifier proved to be quite stable in zero and gain over 0 to 50°C, again more than adequate for the immediate application.

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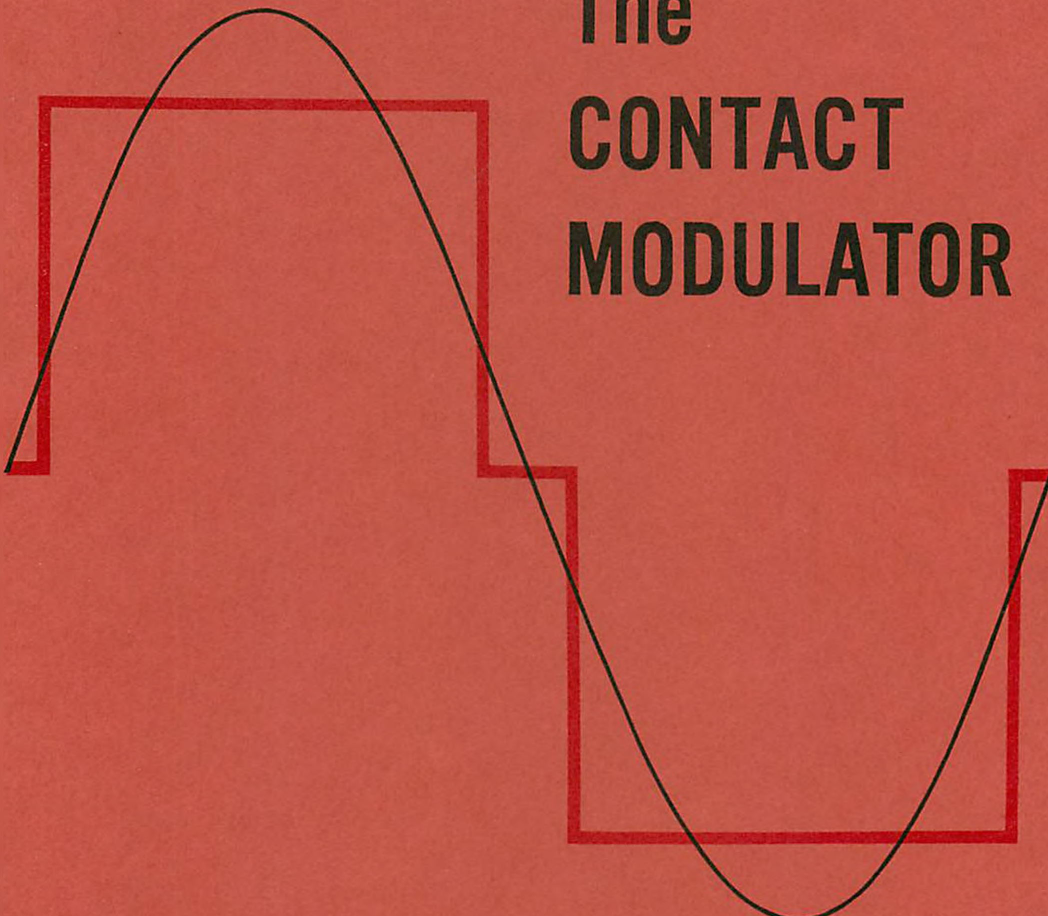

PART 7: Use of Choppers in Systems

THE SEVENTH OF A SERIES

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THE

CONTACT MODULATOR

PART VII

USE OF CHOPPERS IN SYSTEMS

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THE ENGINEERING STAFF

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SECTION I

CHOPPER AMPLIFIER CHARACTERISTICS

In analog computation, the value of performing operations with DC is well known. Even the simple operation of addition is more easily performed if a phase relationship is not involved, and functions such as integration require a mechanical device if the signal is AC.

A simple chopper amplifier such as described in earlier Parts is often used as an operational amplifier, but its response time is limited at least by the chopper frequency and, more often, by filtering employed in the process of DC restoration. This last limitation is not necessarily fatal, as has been outlined in Part VI. If full wave modulation and demodulation are employed, if the chopper dwell time is short, and proper attention is paid to impedance values to reduce charging time in the filter, the response time can be greatly improved. However, computer performance can be considerably improved if the operational amplifiers have response times of better than a millisecond, which is obviously not possible if the carrier frequency is 400 cycles or lower.

Goldberg [1] leaped past the chopper and filter limitations by reducing the chopper amplifier to the function of correcting for error, and the method has become widely used [2] in analog computers built by Reeves, Minneapolis Honeywell, Beckman and others. Response times of 0.25 milliseconds and better are generally used. Since we are now dealing with a DC amplifier of the direct coupled variety, there is no lower limitation on frequency response and by suitable design, the upper limit can be extended to the megacycle region.

With a 60 cycle chopper, the response time is limited to about 30 milliseconds, if the filter does not introduce restrictions, and it is customary to use filter response times of a full second and more with half wave modulation and demodulation in one chopper. It should be noted carefully, however, that it is theoretically possible to obtain a chopper amplifier responding in about ½ a millisecond. This dictates full wave modulation and demodulation, a long dwell time, and a carrier frequency of several thousand cycles. Extension of this response by increase of the chopper frequency still further is a reasonable assumption, but even transistor choppers of the diffused base silicon type have increasing noise,

[1] Stabilization of Wide-Band Amplifiers for Zero and Gain, Goldberg, RCA Review, p. 296, June 1950.

[2] Driftless DC Amplifiers, Bradley and McCoy, Electronics, April 1952.

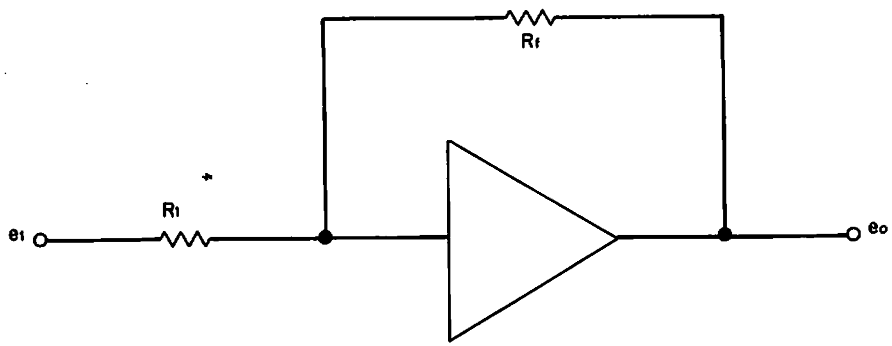


FIGURE 1. *Circuit gain is set by ratio of feedback resistors when amplifier gain is high.*

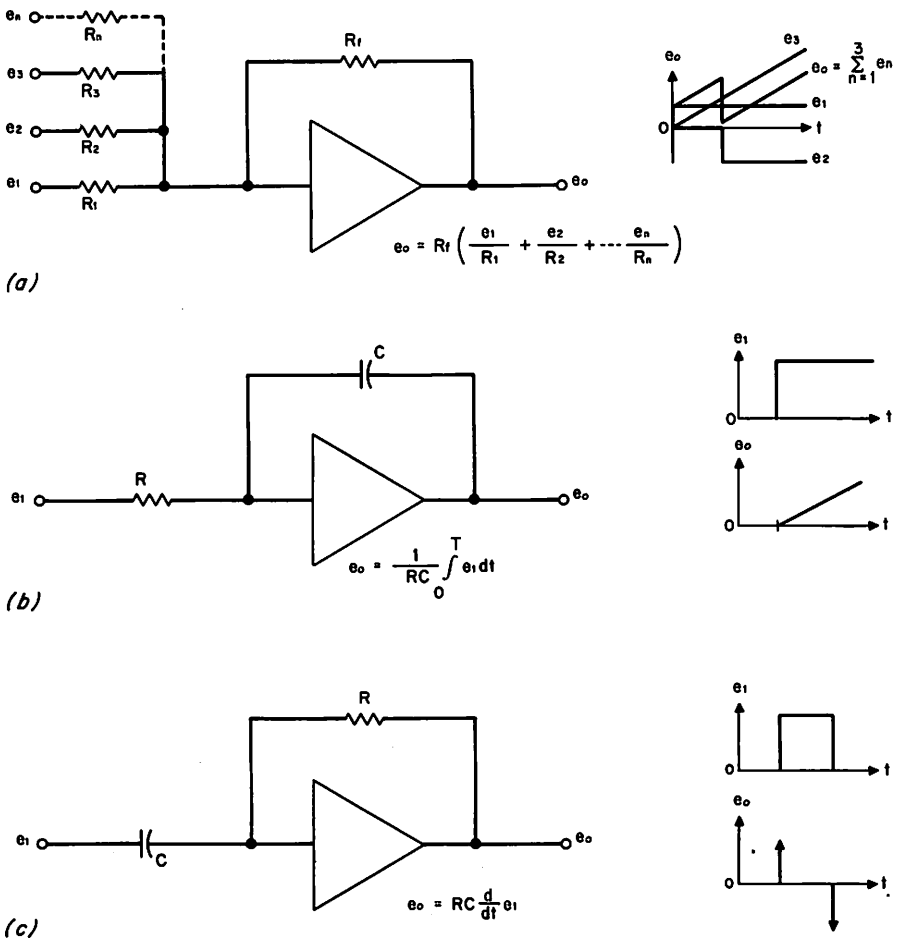
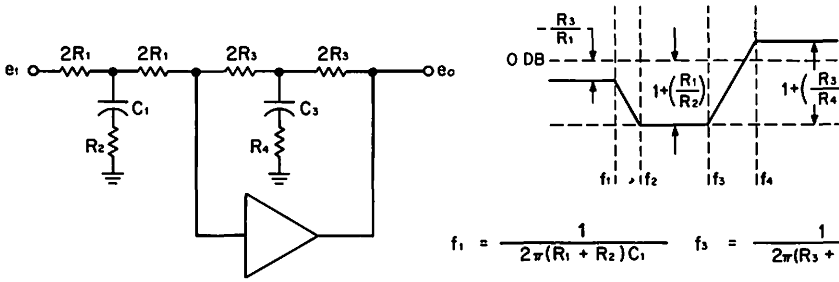


FIGURE 2. *Operational amplifier as (a) adder; (b) integrator; (c) differentiator.*

offset and switching delay, at the higher audio frequencies. If good null performance is also required, i.e., offset of a few microvolts, the limit is quickly reached. Matched silicon transistors are also rather expensive.

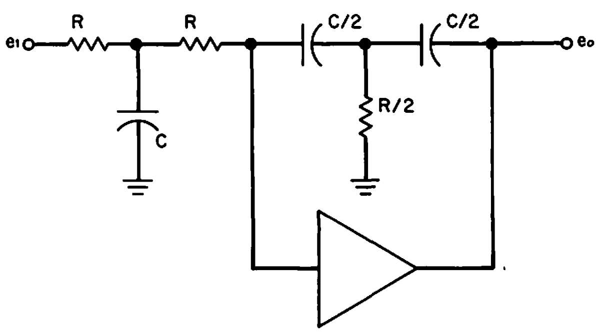
However obtained, a high gain wide band DC amplifier with a stable zero becomes a useful tool in the performance of a great variety of mathematical computations.[3] Circuit gains as astonishingly high as 10 million are reduced to a small value by feedback, as in Figure 1. The system becomes a servo loop summing the input voltage to zero, and



$$f_1 = \frac{1}{2\pi(R_1 + R_2)C_1} \qquad f_3 = \frac{1}{2\pi(R_3 + R_4)C_3}$$

$$f_2 = \frac{1}{2\pi R_2 C_1} \qquad f_4 = \frac{1}{2\pi R_4 C_3}$$

FIGURE 3. *Lead-lag network.*



$$-e_o = \frac{1}{T^2}\iint e_i dt\, dt$$

$$-\frac{e_o}{e_i} = \frac{1}{(T_p)^2}$$

$$T = \frac{RC}{2}$$

FIGURE 4. *Double integrator.*

reducing the input terminal to a very close approximation to ground. Thus the accuracy of the mathematical operations performed can become a function of almost completely passive elements as resistance and capacity, and the accuracy becomes limited principally by the accuracy of these passive elements, Figure 2.

The circuit combinations and functions which can be built upon this foundation will be obvious. Figure 3, describes a lead-lag device sometimes used to compensate for servo system performance, and Figure 4, is an

[3] Electronic Analog Computers, Korn and Korn, McGraw-Hill Book Company.

example of a network combination to obtain a more sophisticated function. Integration is probably the most difficult function performed in analog computation. The accuracy of a chopper stabilized operational amplifier is limited primarily by leakage current in the capacitor and by grid current in the input tube. Both can be made extremely low in good design, and integration accuracies better than 0.1% are practical.

In Figure 5, we have the schematic of a chopper stabilized amplifier. The DC amplifier is composed of an input bidirectional amplifier followed by two more direct coupled stages, the output being a 6AQ5. This requires

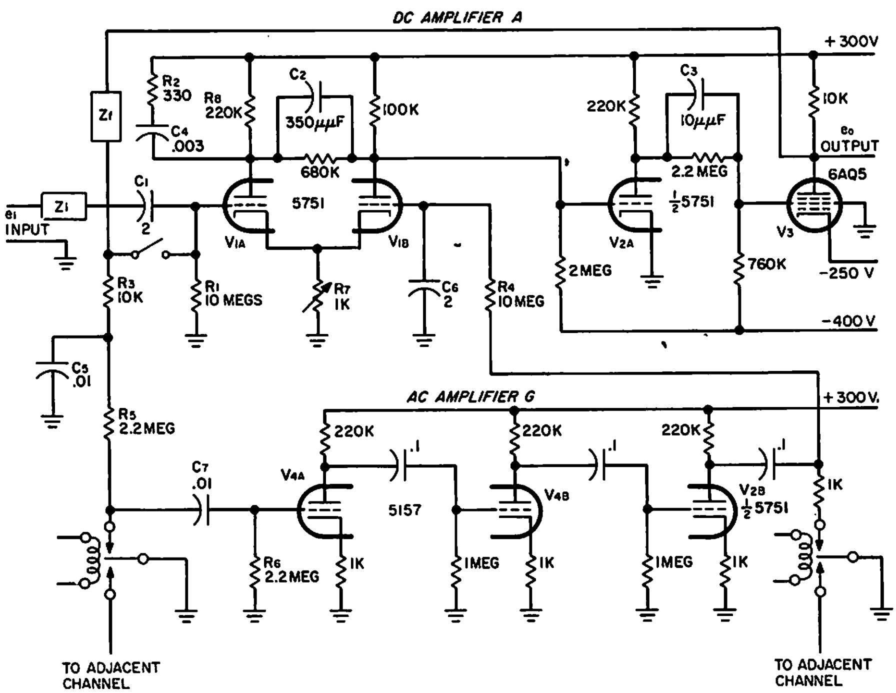


FIGURE 5. *Wide band DC operational amplifier with chopper stabilization.*

the two supply voltages at +300V and −250V. The chopper amplifier contains three *RC* coupled stages. The 5751 is a ruggedized version of the more common 12AX7.

To explain how this circuit utilizes the inherent stability of the chopper amplifier, consider its performance at DC. The capacitor C_1 blocks DC from the first grid of the top amplifier *A*. This connects the two amplifiers, *G* the chopper amplifier, and *A* the DC amplifier, in series beneath the feedback loop consisting of the signal mixing through Z_i and Z_f, which we shall lump into a feedback factor β. Since this is a linear

feedback amplifier, relationships between the various variables are most simply expressed in the language of the signal flow graph.[4] In Figure 6, the through transmission from input e_i to output e_o is the product of the two amplifier gains. The feedback is β. Since A is a DC vacuum tube amplifier, it is susceptible to drift which is referred to its input as an injected noise n.

The output is simply:

$$e_o = \frac{GA}{1\text{-}\beta GA}\, e_i + \frac{A}{1\text{-}\beta GA}\, n = \frac{GA}{1\text{-}\beta GA}\left(e_i + \frac{n}{G}\right) \approx \frac{1}{-\beta}\left(e_i + \frac{n}{G}\right)$$

This expresses the obvious fact that the noise (offset and drift), as it appears in the output, is down by the factor $\frac{1}{G}$ compared to the input signal. This is an echo of the old maxim that noise in the output stage is much less troublesome than noise in the input stage and this is precisely

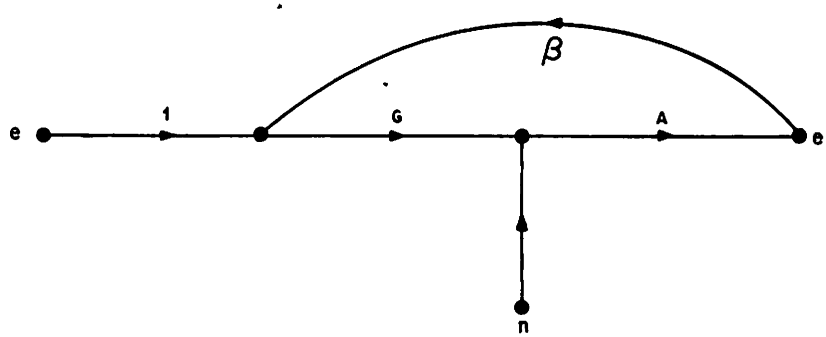


FIGURE 6. *Signal flow graph.*

the reason why chopper stabilization works. The name chopper stabilization seems to imply that the chopper is in some sort of corrective feedback loop which senses the drift at the input and counteracts it. This is a misleading and highly awkward way of looking at the circuit.

Of course, the circuit characterized by Figure 6, would be frequency limited by the low pass band of the chopper amplifier. Now a chopper amplifier as shown in Figure 7, is a sampling data system of sorts. The sampling rate is the chopper frequency in this case, so that the theoretical frequency that can be passed without distortion is one-half the chopper frequency. If one desires to filter the chopper frequency and its harmonics from the output, as we surely do in our case, the pass band is further decreased by the filter R_4 C_6. In our case R_4 C_6 $= 20$, so the upper cutoff frequency of this entire amplifier is about .008 cycles per second.

[4] Feedback Theory — Some Properties of Signal Flow Graphs, S. J. Mason, Proc. I.R.E., September 1953, pp. 1144/1156.

We shall assume that at high frequencies, the chopper amplifier frequency characteristics are determined chiefly by R_4 and C_6.

To utilize the wide band properties of amplifier A (about 10 KC), the amplifier G is simply bypassed as shown in Figure 8. The internal loop gain at frequencies below .008 CPS is GA; above .008 it falls off at −20 DB/decade to A alone until at 10 KC the entire amplifier runs

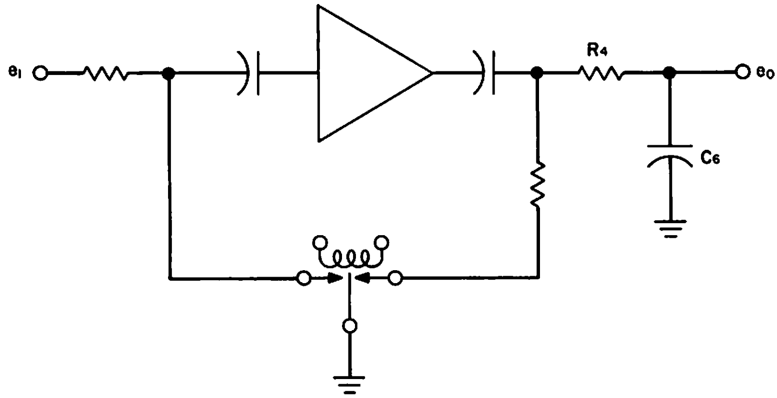


FIGURE 7. *Chopper amplifier is a sampling system.*

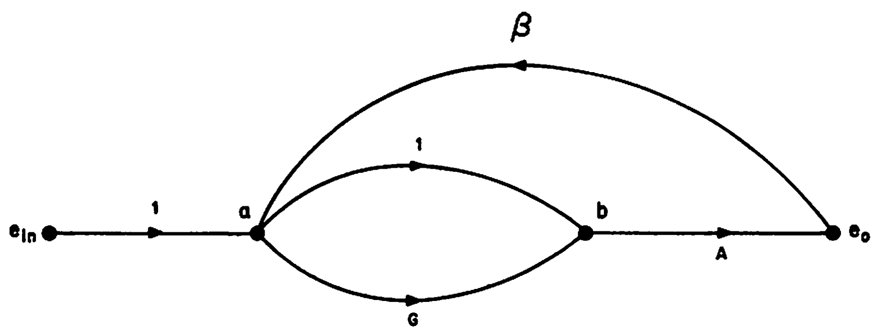


FIGURE 8. *Stabilizing amplifier is bypassed.*

out of bandwidth. Thus the internal loop gain is very much a function of frequency from DC to 10 KC, but in application the return difference βGA is generally so large that overall gain variations are negligible. In this way, wide bandwidth and good stability at DC are combined into one composite amplifier. In the amplifier of Figure 5, the gains A and G are both 80 DB. The gain at DC is thus 160 DB. The bandwidth of A

is about 10 KC. For unity feedback ($\beta = 1$) the overall bandwidth is DC to 100 KC.

Figure 8, contains the feedback loops, βA and βGA. Overall stability is not contingent on the stability of each loop, in fact, if AG is 160 DB, the loop βAG is usually unstable. Therefore, one must consider the graph of Figure 8 in toto, with a feedback loop of $\beta(1 + G)A$. Examination of the Nyquist plot for such a system will disclose that the following is a sufficient but not necessary criterion for stability: (1) the branch G in frequency should behave like a single time constant system whose gain is reduced to zero DB at some frequency (say a decade) less than the -3 DB point of A; (2) the loop βA should of itself be stable. This insures that at frequencies where A begins to introduce phase shift, the contribution at node b from G is less than unity with at least a 45° phase margin and the path G can be neglected.

In Figure 5, the path G has two time constants, $R_3\,C_5$ and $R_4\,C_6$. To satisfy the first part of condition (1), we make $R_3\,C_5 < \frac{R_4\,C_6}{G}$. To satisfy the second part of (1), we observe that at -20 DB per decade, G drops from 80 DB to 0 DB in 4 decades or at a frequency of 80 CPS which is much less than the -3 DB point of A of about 10 KC. One must not ignore the pass band of the chopper amplifier here. It is sufficient if its upper half bandwidth (from chopper frequency to upper 3 DB point) is larger than $\frac{G}{R_4\,C_6}$ (in this case $\frac{500}{2\pi} = 80$ CPS). This is usually not hard to do. To satisfy condition (2) is tantamount to stabilizing the tube amplifier A under unity feedback neglecting G. This is done by the conventional lead networks C_2, 680 K and C_3, 2.2 meg., in combination with the lag network C_4, R_2, R_8, which insures only a -20 DB per decade slope when A has zero DB gain.

The final stability problem may lie in the chopper amplifier G. Figure 7 shows that this RC coupled amplifier has its input and output physically adjacent at the terminals of the chopper tube socket. During any one dwell time, the moving contact is grounding either the input or output which decouples capacitive feedback, but during the off time or transit time, if the chopper is break-before-make, capacitive coupling might cause high frequency oscillations. This will cause a certain amount of drift and offset in the rectified output of G. Most of the capacitive coupling lies in the tube socket and not in the chopper structure. Careful shielding of the socket can reduce this capacitance value to the millipicofarad* region, but experience has shown that when the AC gain of the RC coupled amplifier exceeds about 30,000, off time oscillations become almost impossible to stop. There are two obvious remedies: (a) use a make-before-

*Picofarad — a micromicrofarad, — and our compliments to the editor of the General Radio Experimenter for introduction of this useful European expression.

break chopper or (b) use two physically separated choppers. The former solution assumes the existence of a reliable MBB chopper that will not, in time due to contact wear, convert itself to a BBM chopper.

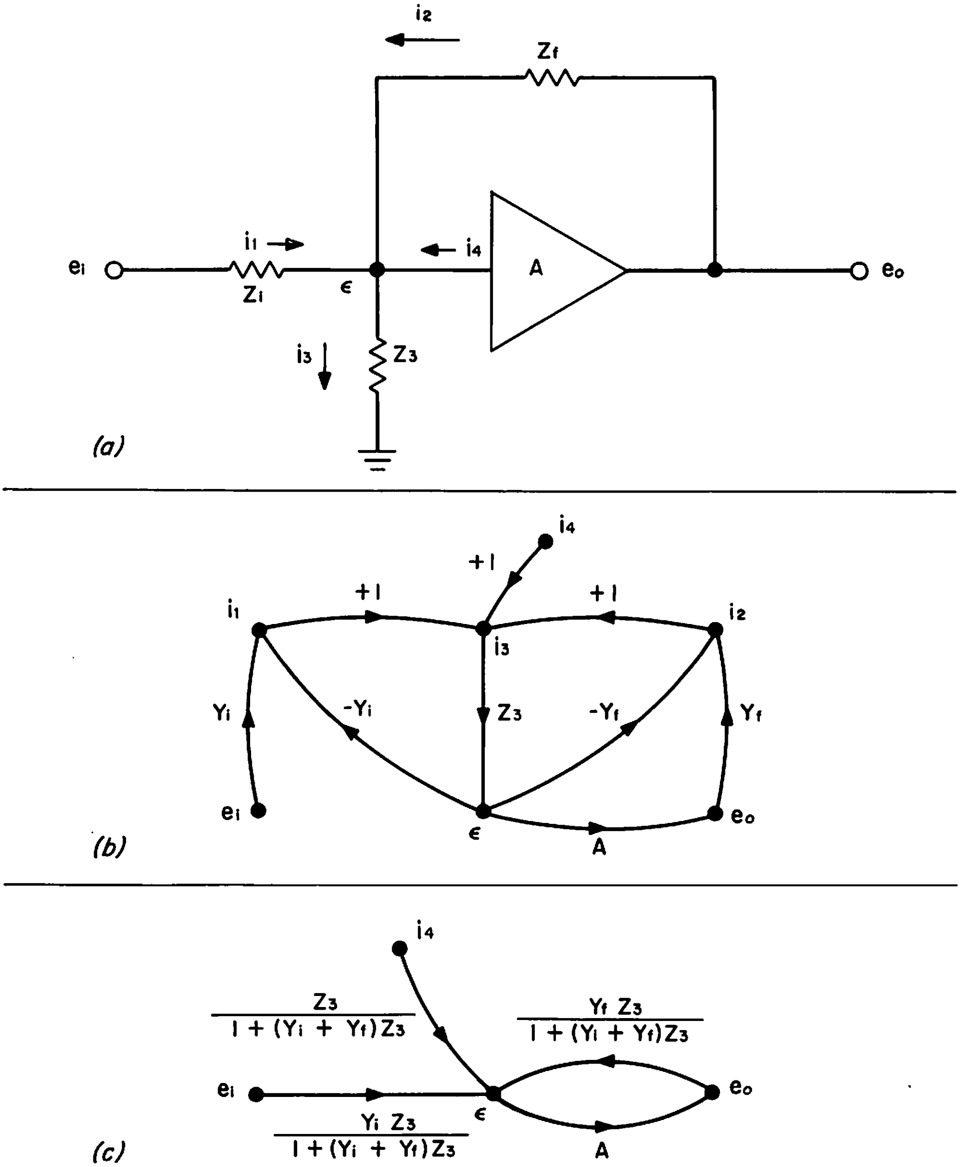
The latter solution seems more reliable and leads to a second choice. With two SPDT choppers available, one can full wave modulate and demodulate the *RC* coupled amplifier and realize efficiency in the form of gain. Another approach is to build two amplifiers, letting one chopper modulate both stabilizing amplifiers and the other demodulate both. Very seldom does an application call for only one operational amplifier, so superfluity is not a problem and in this way the number of choppers per amplifier is reduced to one. In the case at hand, we have chosen this latter course. It should be pointed out that due to the inefficiency of half wave modulating and demodulating, the actual AC gain of the *RC* coupled amplifier must be about 4 times the value of gain desired for *G*, in this case 40,000.

In Figure 9(a), we have added two features for realism, an impedance Z_3 from the summing point to ground and a current flowing out of the amplifier input. The signal flow graph of Figure 9(b), shows the relationships between these variables and 9(c), shows this graph somewhat reduced by elimination of the current nodes. Finally in Figure 9(d), we have the functional dependence of e_o on e_i and i_4 and also Z_3 or A if these values are appreciable. This approximation is excellent if both A and Z_3 are large. If Z_3 is not large, this approximation is still valid if A is exceedingly large to make up for the smallness of Z_3. More rigorously for the approximation to apply, we need $Y_i << (1 - A)Y_f$ and $1 << (1 - A)\ Y_f\ Z_3$. The former implies that $A >> 1$ since Y_i and Y_f may be taken as of the same or adjacent orders. This being the case, the latter is easily satisfied if $Y_f\ Z_3 \geqslant 1$.

The worst case in the amplifier shown is above 1.6 KC where C_5 begins to look like a short circuit leaving $R_3 = 10\text{K}$ to play the role of Z_3. The gain at these frequencies is only 10^4 so that $(1 - A)\ Y_f\ Z_3 = 10^4 \times 10^{-5} \times 10^4 = 10^3 >> 1$, assuming an impedance for Z_f of 100,000 ohms. Although these figures indicate $R_3 = 10{,}000$ ohms is safe, R_3 should probably be increased to at least 100,000 ohms. At DC, Z_3 is about 4.4 megohms, a safe figure especially since A is now 10^8. At any rate, the criterion $AY_f\ Z_3 >> 1$, allows a quantitative lower limit on Z_3.

The second term in the approximation of Figure 9(d) shows the null offset due to currents entering the summing node. Although we have shown this current as injected backwards from the amplifier, i_4 must play the role of <u>all</u> injected currents regardless of their source. Two obvious sources are:

(a) grid current. Some chopper stabilized amplifiers do not use the blocking capacitor C_1 of Figure 5, and overcome this source of trouble by electrometer tubes. This is only a partially successful remedy and



$$e_o = \frac{AZ_3Y_i}{1+Y_iZ_3+(1-A)Y_fZ_3}\, e_i + \frac{AZ_3}{1+Y_iZ_3+(1-A)Y_fZ_3}\, i_4$$

$$\approx -\frac{Z_f}{Z_i}\, e_i - Z_f i_4$$

(d)

FIGURE 9. *Application of a computing amplifier.*

the inclusion of C_1 is a complete cure for very little trouble. C_1 should be of high grade dielectric material, especially with regard to low absorption, such as mylar.

GAIN OF CHOPPER STABILIZING AMPLIFIER G	10,000 AT DC (80 DB)
GAIN OF DC AMPLIFIER A	10,000 (80 DB)
TOTAL INTERNAL LOOPGAIN AT DC	10^8 (160 DB)
DRIFT ATTENUATION AT UNITY GAIN	-75 DB
LOAD	10,000 OHMS
OUTPUT LINEAR RANGE	± 200 V DC
CROSSTALK BETWEEN CHANNELS	-80 DB
NOISE	10 MV WIDE BAND THERMAL
OFFSET AT UNITY GAIN	1 MV

FREQUENCY RESPONSE AND LINEARITY

GAIN	-3 DB BANDWIDTH	PERCENT LINEARITY OVER ±100V RANGE	PERCENT LINEARITY OVER ±200V RANGE
1	100 KC	0.015	0.5
10	30 KC	0.30	1.0
100	20 KC	3.00	10.0

FIGURE 10. *Performance data of dual amplifiers shown in Figure 5 and described in text.*

(b) dielectric leakage across a poor terminal board from adjacent high voltage terminals. Teflon standoffs on a grounded metal terminal board seem an effective cure for this sort of trouble. It should be noted that if Z_f is 10 megohms, a current of 10^{-10} amperes will produce a one millivolt offset referred to the input if Z_i is also 10 megohms.

The subject of chopper noise has already extensively been treated in Part V. We have little to add except that this source of offset strikes at the very heart of the entire amplifier system. Two remarks are important. First, it is clear that to get anywhere at all one must be sure that the circuitry associated with the chopper is neither creating noise nor aggravating the inherent minimal chopper noise (microphonics, cable noise, leaky input capacitor, poor wire placement, creation of flux pickup loops, chassis currents, etc.). Secondly, chopper noise and even associated circuitry noise will cause a null offset at the output, but this is generally a fixed offset susceptible only to changes in the chopper driving voltage. Present day choppers, however, can be obtained with inherent noise levels so

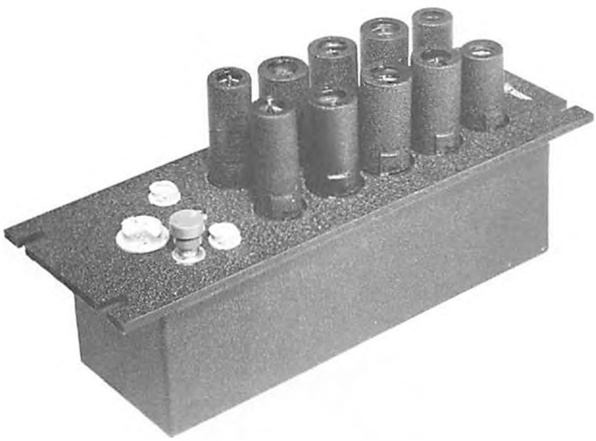


FIGURE 11. *Operational amplifier for analog computer.*

well reduced that the injected offset is of a value comparable with white noise from resistors.

A high *mu* dual triode such as the 12AX7 may, even with cathode degeneration, oscillate when the two half envelopes are in cascade, due to cross talk within the envelope, especially when the low level stage grid is 2 megohms or more above ground. The ruggedized 5751 does not exhibit this annoying habit. Wirewound shielded resistors are helpful in the low level stages of the DC VT amplifier. Deposited carbon resistors sometimes exhibit random low frequency noise fluctuations, making them most undesirable in this part of the circuit. The performance data obtained from the amplifier is shown in Figure 10. The amplifiers were built in pairs, as pictured in Figure 11, thus taking advantage of all chopper contacts.

SECTION II

CHOPPER AMPLIFIER APPLICATIONS

Computer amplifiers similar to the one just described have become available from a number of sources. One such is the Philbrick Type UPA-2, a chopper stabilized operational amplifier shown in Figure 12. Figure 13,

FIGURE 12. *Philbrick Computer Amplifier.*

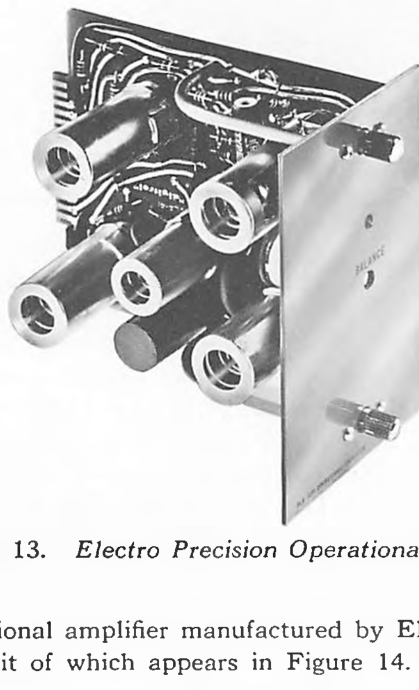


FIGURE 13. *Electro Precision Operational Amplifier.*

pictures an operational amplifier manufactured by Electro Precision Corporation, the circuit of which appears in Figure 14. In this figure, there

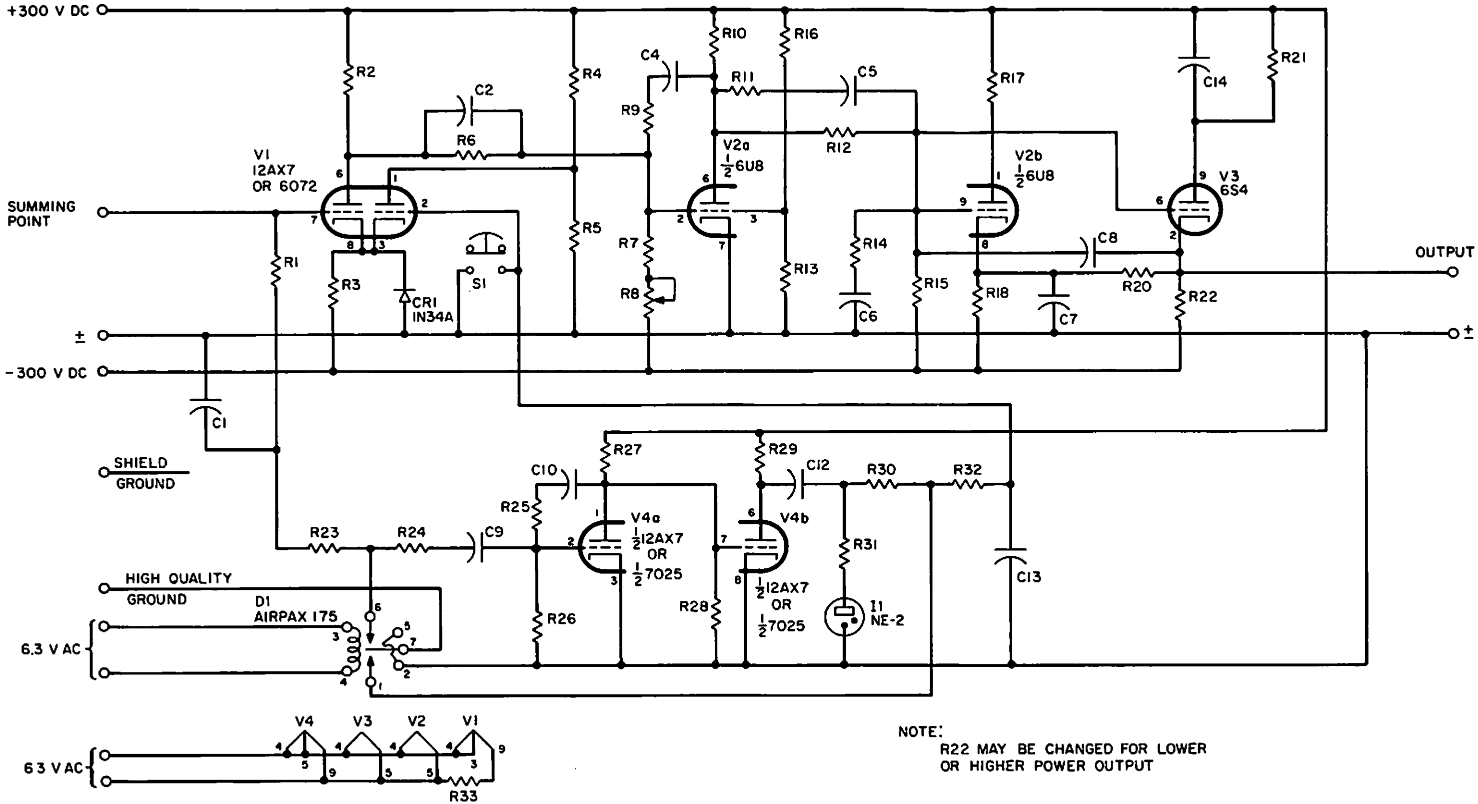


FIGURE 14. *Schematic diagram of Electro Precision Model DLA-41.*

appears a neon bulb I1 and pushbutton S1, a customary method of indicating and quickly relieving overloads which may block the amplifier.

The operational amplifiers just described can be regarded as characteristic of all electronic servo systems, as the effect of the feedback is to sum to zero at the input. A meter inserted as a part of the feedback loop can express the unknown as a function of the feedback current. An instrument described by Leeds and Northrup uses this principal,[5] and is reported to permit current ranges as low as 5 x 10^{-5} microamperes full scale.

Zero stability is obtained by dependence on respectable performance by the chopper, and indeed particular care[6] must be taken if a maximum

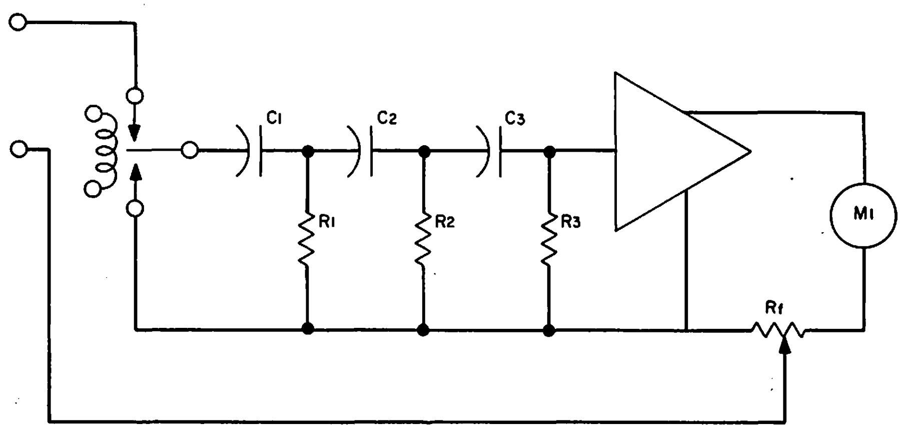


FIGURE 15. *Feedback current can be measured to describe the input.*

of input resistance and sensitivity are needed. The effective input impedance, Figure 15, is considerably increased by the potentiometric arrangement. Range changing can be accomplished by adjustment of the feedback potentiometer R_f, and the desired scales displayed directly on M_1, C_1, C_2, C_3, R_1, R_2, R_3, represent a high pass filter to delete noise created by a slowly varying grid current, an unusual expedient made necessary by the high performance. The use of a low pass filter before the chopper, to reduce noise, and of a high pass filter between chopper and the first stage input, to decouple grid currents or the effect of slow changes in the input tube, was discussed in Part V. As is obvious, these filters will sharply attenuate the frequency response.

[5] Universal Meter for Measuring Voltages at High Impedances, Micromicroamperes, and Insulation Resistance, Clark, Watson and Mergner, Electrical Engineering, p. 41, January 1954.

[6] DC Amplifier With Reduced Zero Off-Set, McAdam, Tarpley & Williams, Electronics, June 1951.

One of the early and frequently used chopper applications lies in zero hunting servos, in which the output of a potentiometer, driven by a motor, is compared to the unknown. Thus we have a mechanical analog of the all-electronic servo. Most frequently a potentiometer is mechanically linked to the parameter to be varied and the system is employed to repeat (or correct) an angle, a distance, a position, a speed, or some other valuable function, as illustrated in Figure 16. The reference voltage is employed as excitation for the potentiometer and the error signal is obtained by comparison at the input chopper. Obviously the stability of the system is, among other factors, a function of the stability of the reference. It is customary to employ a potentiometer having considerable resistance and a long life mercury battery. Note that although the servo sums to zero, if the chopper amplifier input impedance loads the reference potentiometer, an error of position may arise.

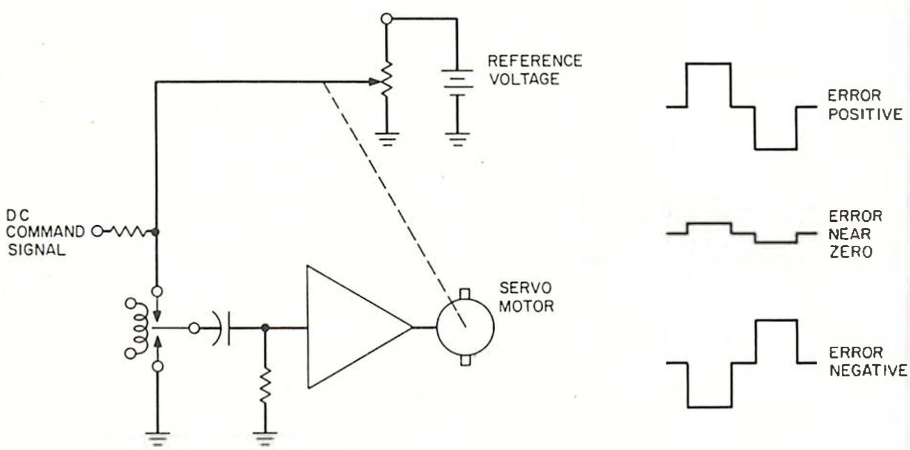


FIGURE 16. *Servo system follows a command signal.*

The "servo motor" operating from the chopper amplifier can conveniently be a phase responsive AC motor, in which the motor provides some degree of power amplification. The amplifier output is used to excite the field, while power is drawn from the line by the armature and delivered to the shaft. In a like manner, the amplifier output can be demodulated and supplied as reversible DC to a motor or to motor control relays.

One way in which substantial power has been handled without the use of a demodulator involves the use of the amplified square wave to fire a thyratron as illustrated in Figure 17. The thyratron firing now permits the switching of a substantial amount of power. Since the anodes of the thyratron must be operated from the same frequency which drives

the chopper, it is obvious that phase adjustments are likely to be involved. Part 4 of this series discusses the question of adjustment of the chopper phase rather extensively.

Figure 18, is the circuit of a chopper amplifier designed for the digital presentation of tachometer information. In this application, the shaft speed is sensed magnetically and delivered to an Airpax Magmeter frequency detector whose output is essentially DC, the average value being a linear function of frequency. The parameter presented to the chopper amplifier input is the integrated average current, and Figure 18, is essentially a milliammeter with digital reading. The chopper compares the current input with the 1,000 ohm potentiometer driven by the servo

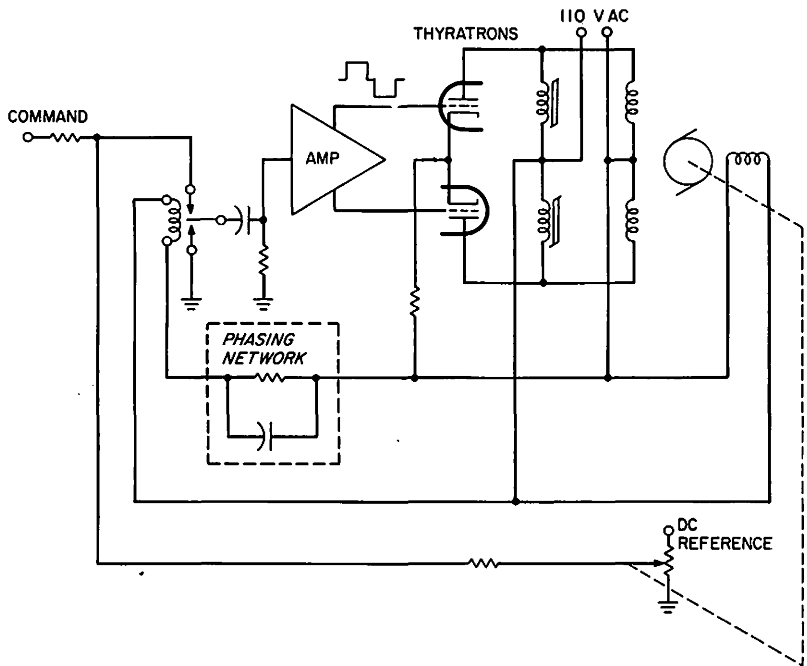


FIGURE 17. *Square wave from chopper triggers thyratron tubes.*

motor. The digital presentation is a mechanical turns counter driven by the motor.

The chopper output is, of course, a square wave whose phase and amplitude are a function of the polarity and amplitude of the different signals appearing at the chopper output. The 10,000 ohm resistor in series with the base of the first transistor provides limitation of the relatively high signal level which can occur at a distance from the null. A complete tachometer employing this circuit appears in Figure 19.

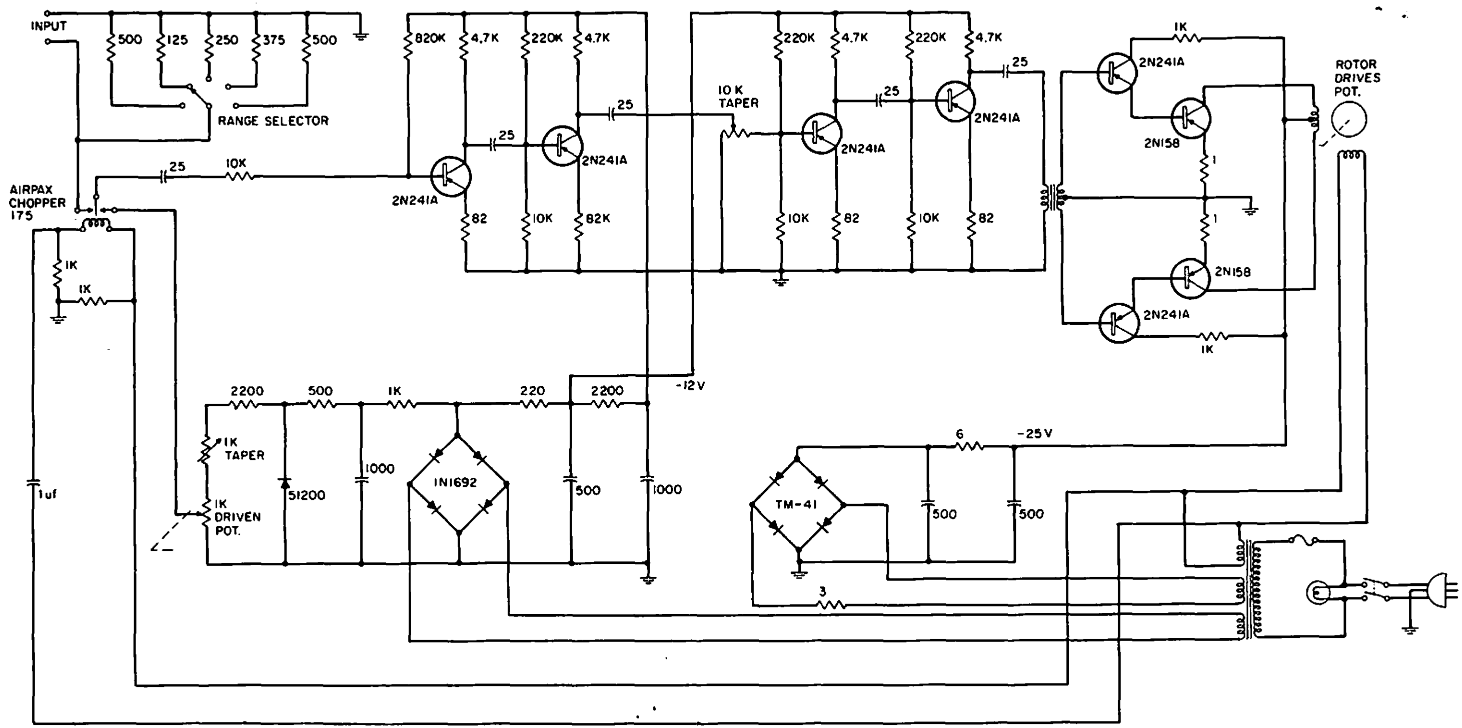


FIGURE 18. *Electronic circuit for digital presentation of current values.*

The basis of almost all precision measurement lies in observing the difference between a standard and the unknown quantity. The precision of the potentiometer method of voltage measurement is well known. Extension of this accuracy and a reduction in loading becomes possible by periodic

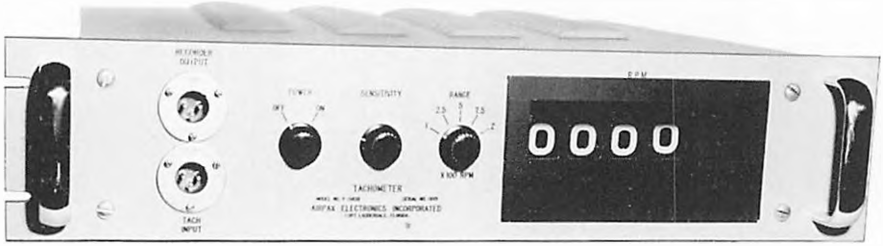


FIGURE 19. *Airpax Digital Tachometer.*

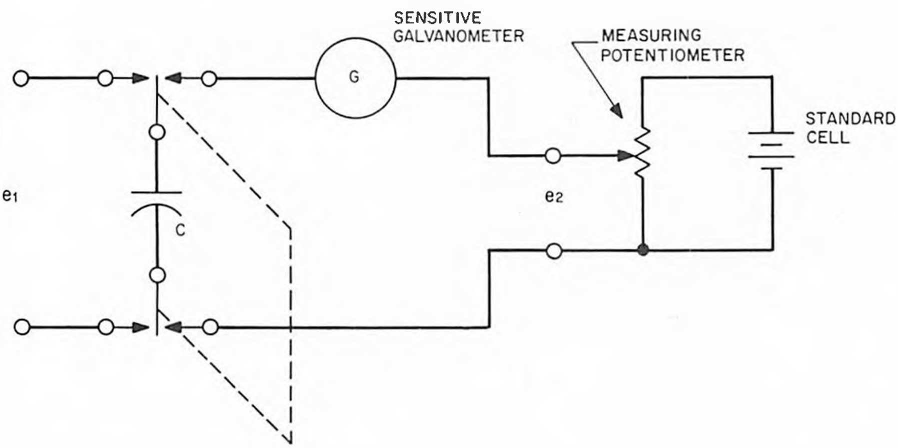


FIGURE 20. *Sampling circuit measures e_1 by use of DPDT chopper.*

sampling,[7, 8] Figure 20. The charge on capacitor *C* is compared with the potentiometer output e_2 by adjusting for zero galvanometer deflection. Since the error signal at *G* is already a chopper output, the application here of an AC amplifier is obvious, if we can rapidly move the switch back and forth.

[7] An Isolating Potential Comparator, Dauphinee, Canadian Journal of Physics, Vol. 31, p. 577, 1953.

[8] Dispositif Electronique Pour La Mesure Precise Des Tensions Continues Par Comparaison, Aumont and Romand, Revue Generale De L'Electricite, April 1953.

Chopper amplifiers have been of particular use in digital voltmeters, because of the pressure naturally arising for maximum gain, freedom from drift, good linearity and fast response. This kind of performance is most readily obtained with a mechanical chopper. Other modulation methods can approach this performance, but none can surpass it. In company with a vacuum tube input, it becomes possible to obtain high impedances in addition.

There are two general types of digital voltmeters, both of which employ the sampling technique. One is a direct reading device employing a zero hunting servo, and the other a manual adjustment of the potentiometer, employing many decades. This second device is also a zero hunting servo, with the gain of the technician interposed in the feedback loop. It is notable that the technician as an amplifier has an excellent signal to noise ratio and while occasionally considered unreliable (like choppers), has a tendency to outlast the rest of the gear (like choppers).

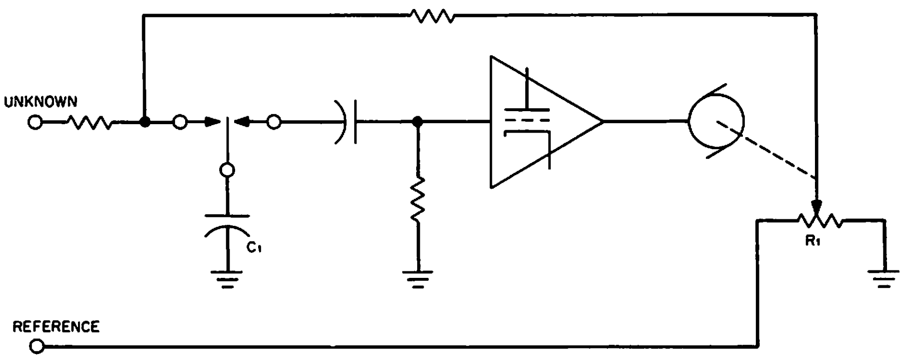


FIGURE 21. *Chopper amplifier hunts for zero.*

Figure 21, shows one common type of zero hunting servo amplifier. The chopper performs a sampling function and delivers the sample to the amplifier; in this illustration the sample is the difference voltage existing between the unknown and the reference. Please note an additional factor. Capacitor C_1 is essentially an integrating device, and delivers to the amplifier the average difference value. Thus it tends to delete noise spikes in potentiometer R_1. The motor at the amplifier output is so phased that it drives toward zero signal as summed in the chopper input.

Several devices, [9, 10, 11] have been built using this principle such as the digital voltmeter of Hycon Manufacturing Company, 1030 South

[9] Design of a Ratiometer, Kuehn, Electronic Equipment, November 1955.

[10] Digital Presentation Vacuum-Tube Voltmeter, Nuut & Munsey, Electronics, January 1956.

[11] Digital Voltmeters, Editor, Electromechanical Design, June 1958.

Arroyo Parkway, Pasadena, California, illustrated in Figure 22. In these devices, a ratio is established against a standard reference such as a mercury cell, additional voltage ranges are obtained by divider networks, and rectifiers are used to provide AC ranges. The Hycon Model 615 compares a portion of the voltage under test to the potential of a mercury reference cell. A chopper is used to alternately sample the two voltages. The output of the chopper comparator is amplified and used to energize the control winding of a servo motor which drives the three cylindrical dials to provide the digital readout.

The Fluke* Model 801, Figure 23, potentiometric DC voltmeter introduces the operating technician in the feedback loop and thereby enjoys a claim to an accuracy of 0.05% over a range of 0.1 volt to 500 volts. A standard cell is used as a reference, calibrating a well regulated voltage. This voltage is supplied to a precision six-element decade potentiometer

FIGURE 22.
Hycon Digital Voltmeter.

FIGURE 23. *Fluke Model 801 Potentiometric Voltmeter.*

which is then nulled against the unknown. A chopper amplifier, using a 60 cycle Type 175 Airpax chopper, provides sensitivity to permit accuracy at low voltages.

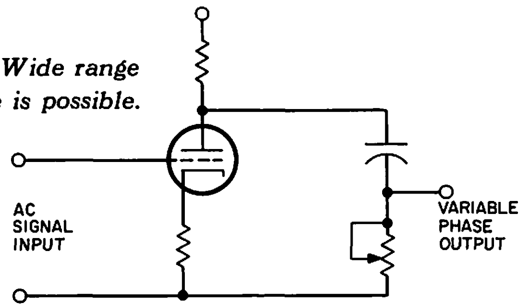
It is not always possible to use an identical chopper for demodulation. This condition is likely to arise when substantial power must be handled at the output. Airpax choppers operate successfully with comparatively large values of current and voltage, as well as minute values. However, currents in excess of 1 or 2 milliamperes and voltages in excess of 100 volts, very frequently require that we give special attention to the internal design. The design problem confronting the chopper engineer then begins to look more like a power vibrator application.

* John Fluke Manufacturing Co., 1111 W. Nickerson Street, Seattle, Washington.

Control of the phase angle of the chopper is frequently necessary, and the phase stability can be arranged to become comparatively good over a normal range of generator frequency such as 380-420 cycles. When the phase angle of the signal is to be varied, Figure 24 illustrates one method in which a range of 0 to 180° is easily obtained.

When the chopper signal is used directly, as in Figure 17, the square wave output is usually satisfactory. In fact, this form is often necessary to prevent loss of information. But if it were possible to obtain a sine

FIGURE 24. *Wide range of phase angle is possible.*



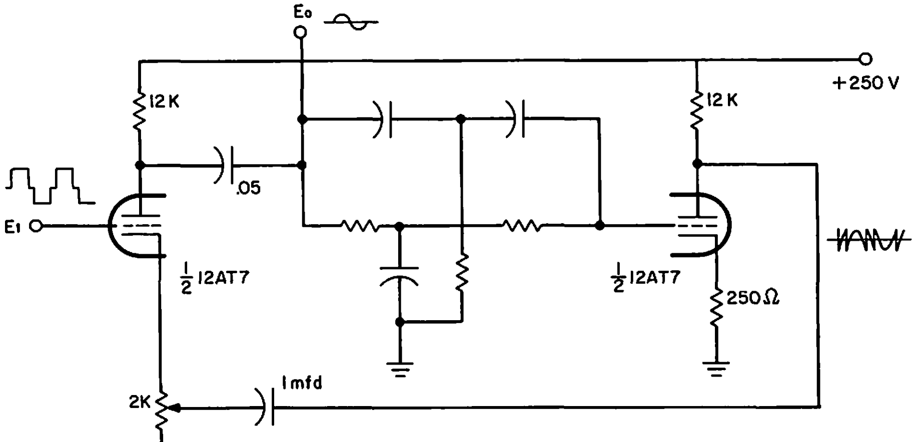


FIGURE 25. *Filter circuit delivers sine wave.*

wave from the chopper contacts, again referring to Figure 17, the firing angle of the thyratrons could be delayed. In this circuit, of course, they fire when the chopper contacts close unless methods are used to introduce delay. Other devices which prefer a sine wave will occur to the reader.

A band pass filter can, of course, be used to reject the harmonic content. One of the penalties paid when a narrow band amplifier is used is a severe reduction in response time. (There are also advantages, such as an improved signal to noise ratio.) A simple and effective filter circuit is described in Figure 25, which sidesteps the response time limita-

tion. A twin T filter between stages rejects the fundamental signal. Feedback is clamped around the amplifier. Since the notch filter has rejected the fundamental, the negative feedback is composed of noise and harmonics, resulting in an excellent sine wave output whose phase angle is not different from the original square wave.

While chopper wave forms are sometimes filtered, there are also applications where the chopper is described as a filter.[12] The output signal from an element, such as a resolver, will be AC but may contain noise capable of disturbing system performance. Filtering is, of course, easier if the signal can be DC. One method of operating on the noise is to demodulate, filter, and then remodulate, Figure 26, keeping the desired information without the noise. This procedure can also be viewed as the process of integrating the random variations, thus delivering information of perhaps greater accuracy.

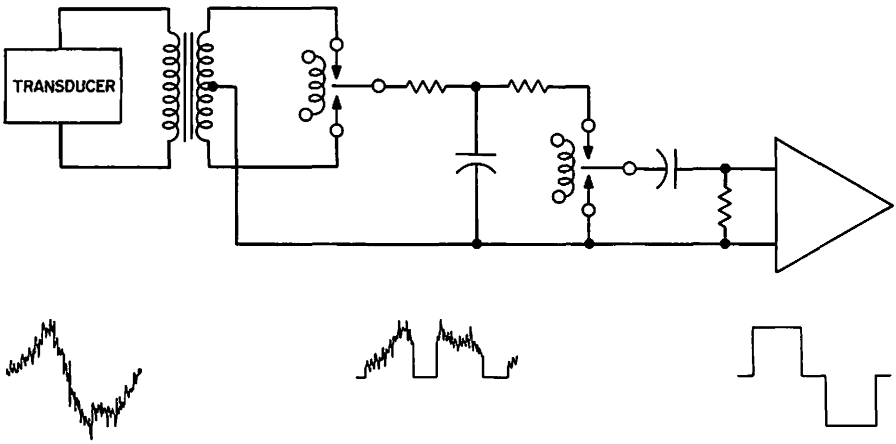


FIGURE 26. *Reduction of signal to DC permits operating on random variations or noise.*

Used as a demodulator, a chopper provides an offset free method [13, 14] of obtaining random functions, or "white" noise, at low frequencies, down to zero frequency. This accomplishment is described in Figure 27. The random noise generated in a gas tube is amplified and limited in its average value by a regulating circuit. This output of the amplifier contains noise components whose spectrum is limited principally by the amplifier bandwidth. A filter follows, limiting the bandwidth to about twice the desired

[12] Analysis of a Special Purpose RC Filter Incorporating a Periodically Conducting Bilinear Element, Bolie, Proceedings IRE, p. 1435, September 1954.

[13] Stabilized Noise Source for Air Weapons Design, Beecher, Bennett and Low, p. 163, Electronics, July 1954.

[14] Use of Noise and Statistical Techniques in Analog Computation, Low, Chapter 26 of Vol. 2, Handbook of Automation, Computation and Control, Wiley & Sons.

noise spectrum and delivers this limited spectrum to a chopper demodulator. The random signal thus obtained has a variety of interesting uses, such as the representation of turbulence, computation of air foil stresses, and the development of probable courses of action in war games.

Figure 28, describing the performance of the filters and the spectral density of the final signal delivered, is particularly helpful in illustrating

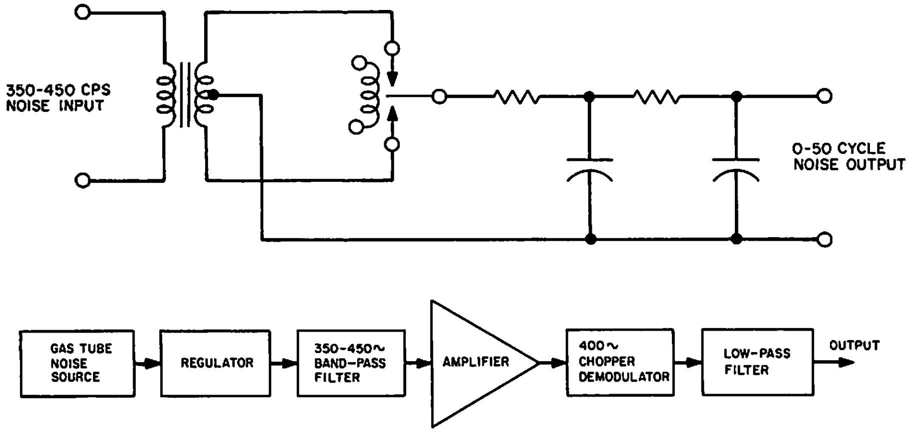


FIGURE 27. *Chopper used as detector for a noise generator.*

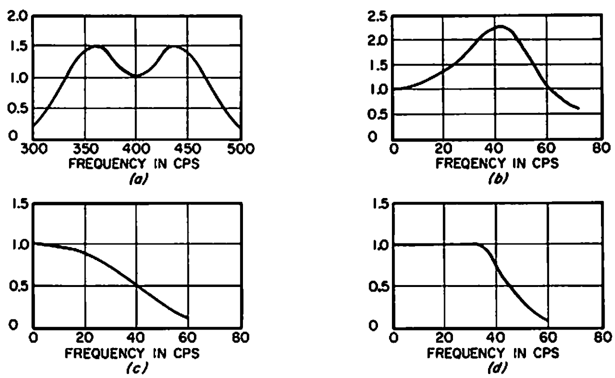


FIGURE 28. *Filter response and noise spectral density show output spectrum development.*

the principle. A demodulator is a duality, being a modulator when viewed differently. In this application, the chopper can be considered a mixing device, since the output is essentially the sum and difference of noise signal plus the heterodyned frequency of 400 cycles. Since the random noise delivered at the filter output covers the spectrum of 350 to 450 cycles, as shown in Figure 28, the output will be 0 to 50 cycles. A low

pass output filter further attenuates higher frequency components. Figure 29, pictures a noise generator employing this principle, built by Elgenco*, and another offered by Electronic Associates**, appears in Figure 30.

The difficulty of precisely measuring a difference frequency of a few cycles or less is obvious unless counting techniques and a fairly long time scale are used. If stability in the vicinity of 1 part in 100 million is desired, the problem is obviously more difficult. Automatic control of frequency by counting methods becomes quite complex. A precision frequency generator ascribed to P. G. Sulzer [15] is illustrated in Figure 31. A 60 cycle chopper is used to modulate the output of a bridge in which the crystal functions as one series resonant arm. The modulator unbalances the bridge on each half cycle, delivering a signal to the bridge detector, in this case a high gain RF amplifier. A second chopper functions as a de-

FIGURE 29. *Elgenco Model 301 Noise Generator.*

FIGURE 30. *Electronic Associates Model 201A Noise Generator.*

modulator and drives a reactance tube, which in turn controls the oscillator to the crystal frequency. The circuit is reported to provide stability of 1 place in 10^9 even over long time periods.

Airpax chopper Type 199, shown in Figures 32 and 33, is a "coaxial" chopper designed to have a standing wave ratio of less than 1.2 over a frequency range of 100 to 400 megacycles. The connections to the cavities are made with Microdot coaxial connectors, and the SWR figure given above is in terms of the chopper assembly with the necessary interconnections.

[15] High Stability Bridge — Balancing Oscillator, Sulzer, p. 701, June 1955, Proceedings IRE.

* Elgenco, Inc., P. O. Box 90344, Airport Station, Los Angeles 45, California.

** Electronic Associates, Inc., Long Branch, New Jersey.

The driving frequency is normally 100 cycles. Type 199 is designed to have mechanical resonance at somewhat over 200 cycles. This permits the moving armature to ignore harmonics of the driving signal, and thus operate successfully from the square wave output of a vibrator or transistor inverter. The effect of using a high resonant frequency from a highly

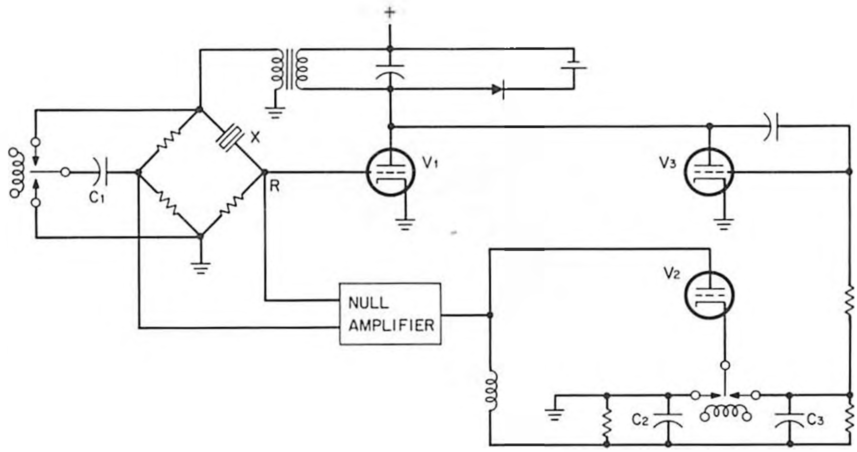


FIGURE 31. *Simplified schematic of bridge-balancing oscillator.*

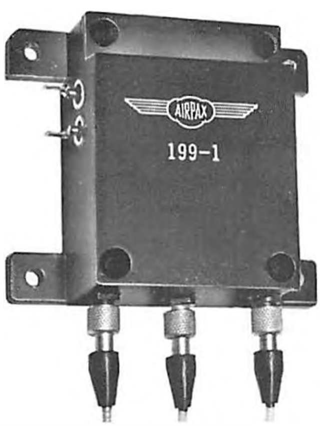


FIGURE 32. *Airpax Type 199 coaxial chopper.*

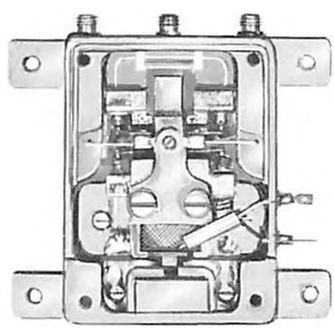


FIGURE 33. *Internal construction of Type 199 chopper.*

distorted driving source is perhaps intuitively obvious — a moving arm having a high response speed is free to follow the source. The result is likely to be unbalance of dwell time (Part II) or apparent erratic contact action.

The Collins Radio Company manufactures UHF direction finding equipment, employing the Type 199. Figure 34, describes one of their products, the ARA-48, used principally in military aircraft. The chopper is used to modulate the signal picked up by a rhombic antenna in a somewhat similar manner to the ADF receivers used at frequencies up to 1000 KC. The rhombic antenna is driven to a null between the forward

FIGURE 34. *Collins ARA-48 UHF direction finding equipment employs Airpax 199 chopper.*

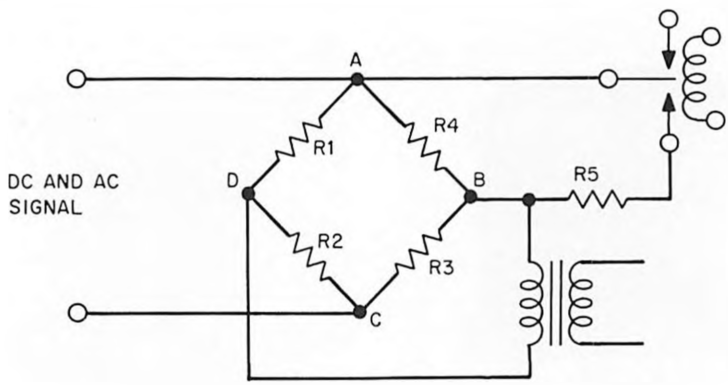


FIGURE 35. *Modulator circuit permits wide frequency response.*

and back cardioid patterns. A terminated rhombic antenna has a unidirectional cardioid pattern, and the double pole double throw chopper contacts switch the input to the VHF receiver, and also the terminating resistor from front to back of the antenna.

A method of extending the possible response time of a chopper amplifier is reported by Raymond Bark of Boeing Airplane Company. In Figure 35, assume as one suitable arrangement, that $R_1 = R_2 = R_3$,

but that R_4 is higher. R_5 is of such a value that the parallel combination $\frac{R_4 R_5}{R_4+R_5}$ is less than R_1, or R_2, or R_3. The signal appearing at the output of the bridge, terminals *BD*, will then reverse in polarity depending on whether the chopper contacts are closed or open.

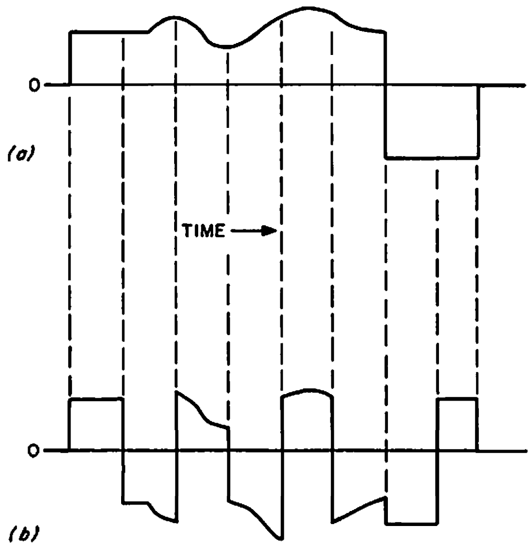


FIGURE 36. *Information is not lost during chopper transit time.*

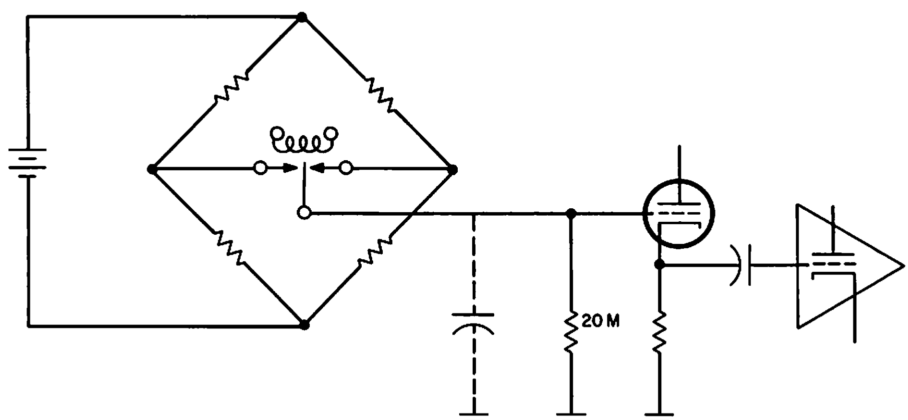


FIGURE 37. *Choppers make good bridge detectors.*

The advantage is that information contained during chopper transit or off time will not be lost. This effect is pictured in Figure 36. Experimental response of several thousand cycles is practical. There are attendant

beats and ambiguities in the vicinity of the coil excitation frequency and harmonics.

As a detector for DC bridges the chopper has a number of distinct advantages, in particular for precision balancing at high impedances. In Figure 37, the chopper samples both sides of the bridge. The extremely long time constant of the input capacity and insulation resistance prevents the input tube from recognizing the chopper off time. The capacitor shown from grid to ground is usually unnecessary, particularly as it is often necessary to shield the grid connection with consequent high input capacity. In the off-balance condition, the chopper delivers an AC wave whose phase and voltage are a function of the polarity and voltage of the unbalanced output. An ordinary rectifier operating a DC meter provides an extremely sensitive bridge indicator capable of unusually precise null adjustment.

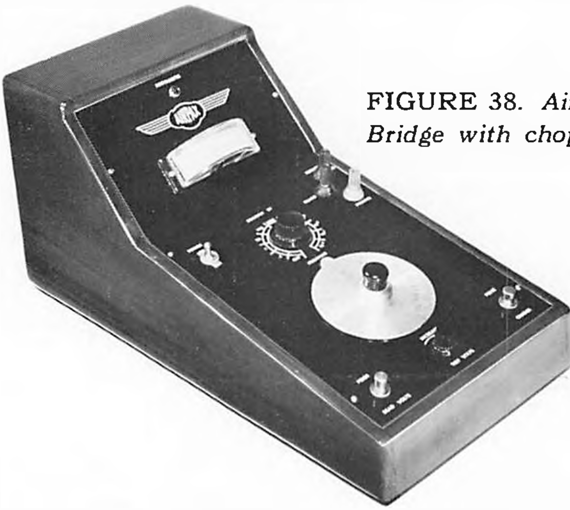


FIGURE 38. *Airpax Megohm Bridge with chopper detector.*

Page 12 of Part IV describes a capacitor modulator useful in reading small voltages at high impedances. In a contact making chopper, the grid current of an input tube may be modulated and if so, will become a source of noise and offset. However, the effect can be minimized by the use of electrometer tubes, which are designed for extremely low grid current, (reaching values as low as 10 — 15 micromicroamperes), and it becomes possible to delete the input capacitor.

An experimental megohm bridge developed by Airpax is illustrated in Figures 38 and 39. The need for this bridge arose originally from the problem of accurately measuring the back resistance of diodes at less than the zener voltage. The bridge has a range from 0.1 to 10^{11} megohms, an accuracy of 3% up to 10^5 megohms, and 5% up to 10^9 megohms. Voltage across the unknown may be adjusted from 25 to 500 volts. Detection is accomplished by sampling the bridge output with a chopper and amplifying, as shown in Figure 39. The diode section of the 6AV6 final

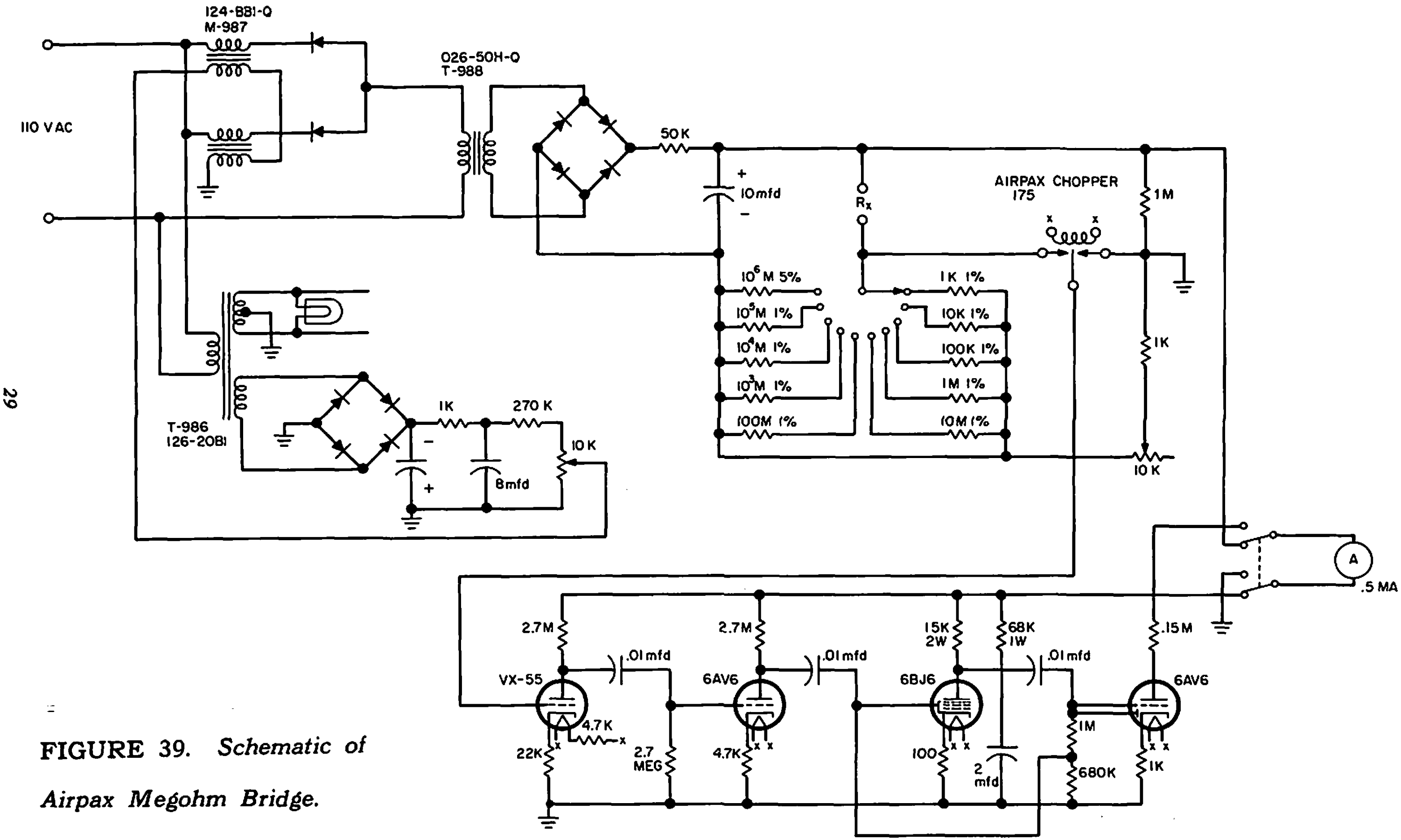


FIGURE 39. *Schematic of Airpax Megohm Bridge.*

stage peak rectifies the amplified signal, biasing the grid of the final. Indication is provided by a 500 microampere meter, which is also used to read voltage. At zero signal the meter is at full scale and at lesser values with signal, thus it is obviously impossible to damage the movement by overload. Stable bridge voltage is supplied by the magnetically regulated power supply. As mentioned, a VX-55 electrometer tube is used at the input to eliminate the input capacitor and permit direct coupling, possible because of the extremely low grid current of the VX-55.

Another use to which choppers can be put is that of component resolving. Two choppers running 90° out of phase are used to demodulate a carrier signal. Since the chopper is a phase demodulator, the DC com-

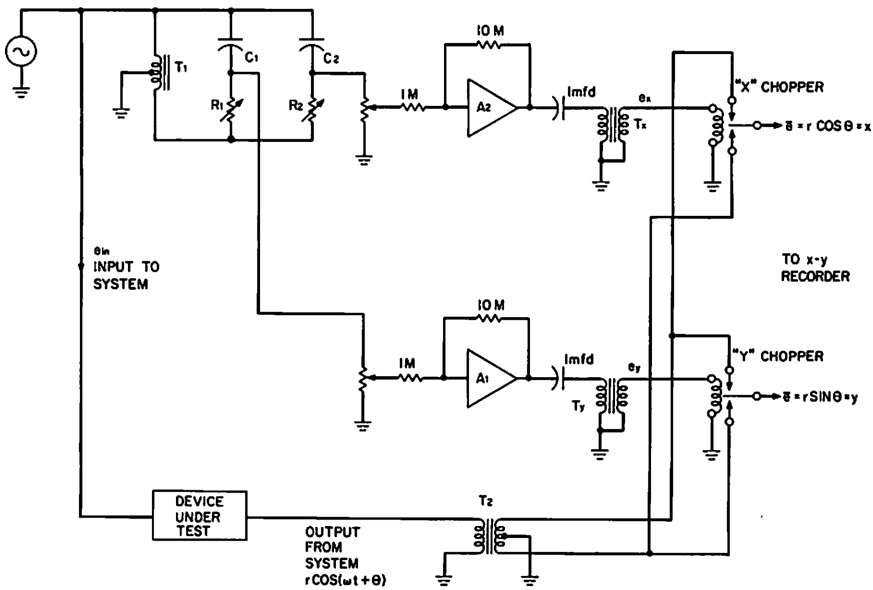


FIGURE 40. *"X" chopper is driven in phase with e_{in} while the "Y" chopper lags by* $\frac{\pi}{2}$.

ponent in the output of one of the choppers is *r sin θ,* while the DC component in the output of the quadrature chopper is *r cos θ,* where *r* is the amplitude of the carrier wave and *θ* is its phase angle with respect to some arbitrary chosen reference phase. Such a system then takes a carrier wave of the form $r \sin(\omega t + \theta)$ and breaks it down into its components. By applying these two components to the *x* and *y* inputs of either an oscilloscope or a pen recorder, it is possible to represent the input wave as a vector or a phasor on a phase plane. In this way, it is possible to draw Nyquist diagrams directly. Figure 40, shows such a component resolver

being employed in conjunction with an analog computer to study the transient behavior of non-linear tuned filters.

Phasing network C_1, R_1, provides a signal to amplifier A_1 which drives the y component chopper. Similarly, phasing network C_2, R_2, provides a signal through amplifier A_2 to drive the x component chopper. The phasing networks compensate for the phase angles of the choppers themselves as well as assuring a 90° phase difference between the two choppers. Thus the x chopper is running in phase with the system input voltage while the y chopper is leading the system input voltage by 90°. The system input voltage is fed into the system under study.

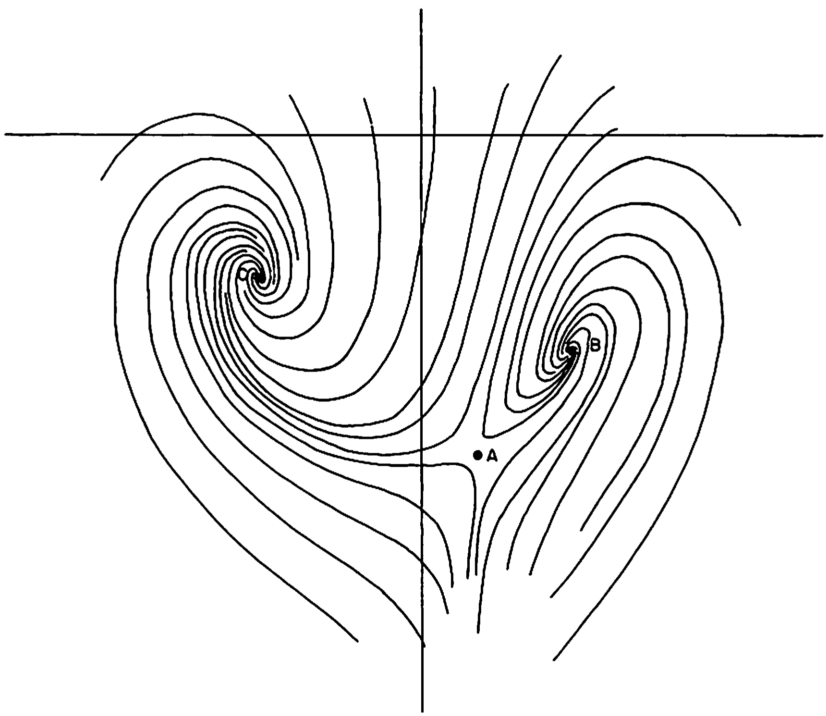
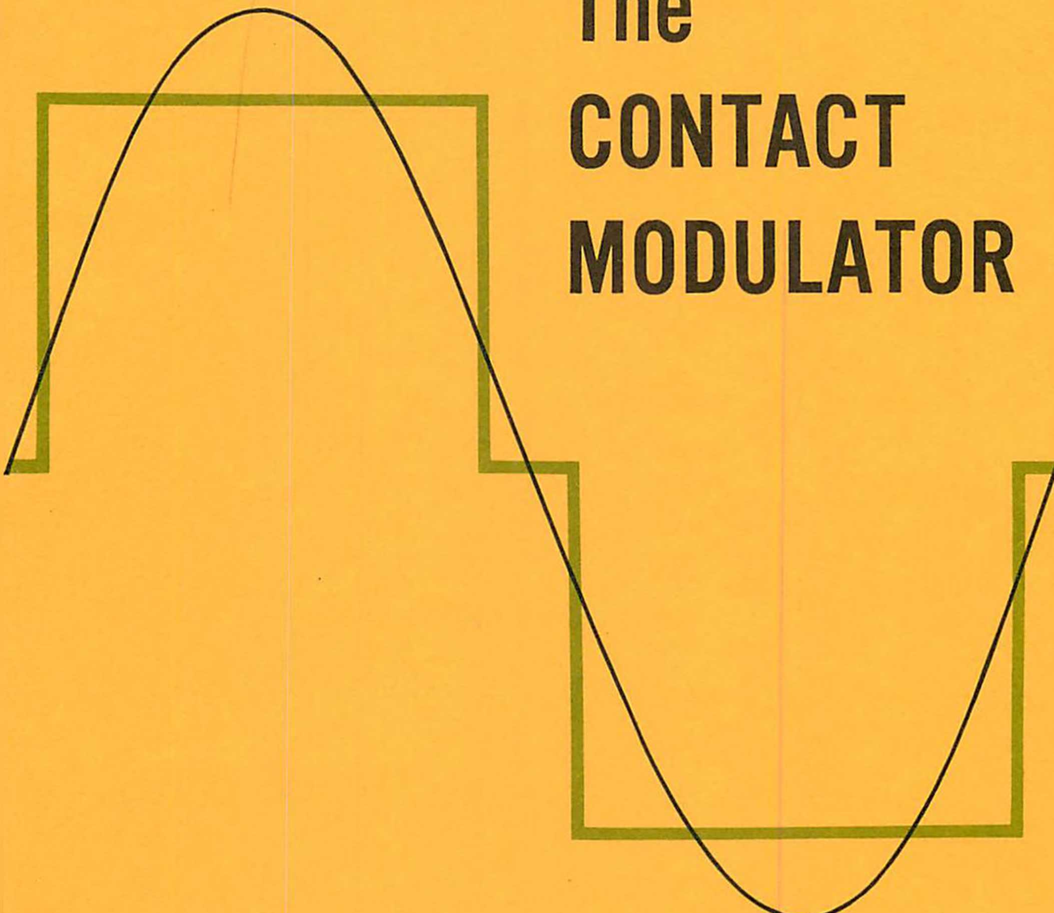


FIGURE 41. *Phase planes plotted with circuit of Figure 40.*

In this particular example, the component resolver drives a second order non-linear differential equation synthesized on an analog computer. The output of the system under study was a voltage of the same frequency as the input, but contained both amplitude and phase modulation. If the system were shock excited the xy recorder would trace out a phase plane trajectory indicating the system's dynamic response directly on the phase plane. By varying the initial conditions, it is possible to generate an entire family of such trajectory as shown in Figure 41.

AIRPAX

The CONTACT MODULATOR



PART 8: The Literature and History of Choppers with Catalogue Section

THE EIGHTH OF A SERIES

AIRPAX

THE
CONTACT MODULATOR

PART VIII

THE LITERATURE AND HISTORY OF CHOPPERS WITH CATALOGUE SECTION

WRITTEN BY

THE ENGINEERING STAFF

OF

AIRPAX ELECTRONICS

INCORPORATED

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SECTION I

THE LITERATURE AND HISTORY OF CHOPPERS

The patent office is a fairly accurate source of dates. One does have grave difficulty deciding just what was invented (from claims). However, the diagrams and explanations mirror the time, the art, and its language. And, the patentor will probably receive the credit for the contribution to progress, be it deserved or not.

The science of using a carrier system to convey information has a long history. A contact-making device as a modulator is unique in that

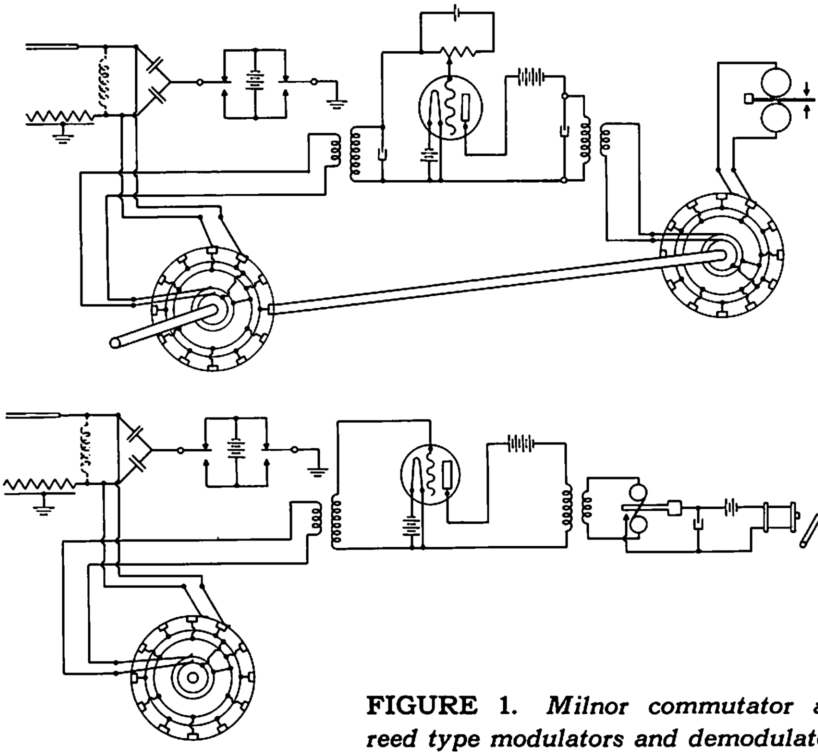


FIGURE 1. *Milnor commutator and reed type modulators and demodulators.*

its conducting phase approaches perfect conduction and its non-conducting phase represents a reasonable approximation to an infinitely high resistance.

It is therefore not too surprising to find that in 1918 one J. W. Milnor,[1] concerned with the problem of weak signals from the submarine telegraph, successfully employed contact modulators and demodulators to amplify the low level (and low frequency) telegraph signals. Milnor naturally tried rotary commutation, however, he also describes a tuned reed contact device,

[1] Amplifier, J. W. Milnor, U.S. Patent No. 1,378,712, filed January 17, 1918.

the use of transformers, and disk rectifiers. Figure 1, reproduces some of the descriptive matter in the Milnor patent.

Much later, in 1939, Eberhardt [2] described a variety of contact and other modulation devices and outlined their uses. Some of his patent

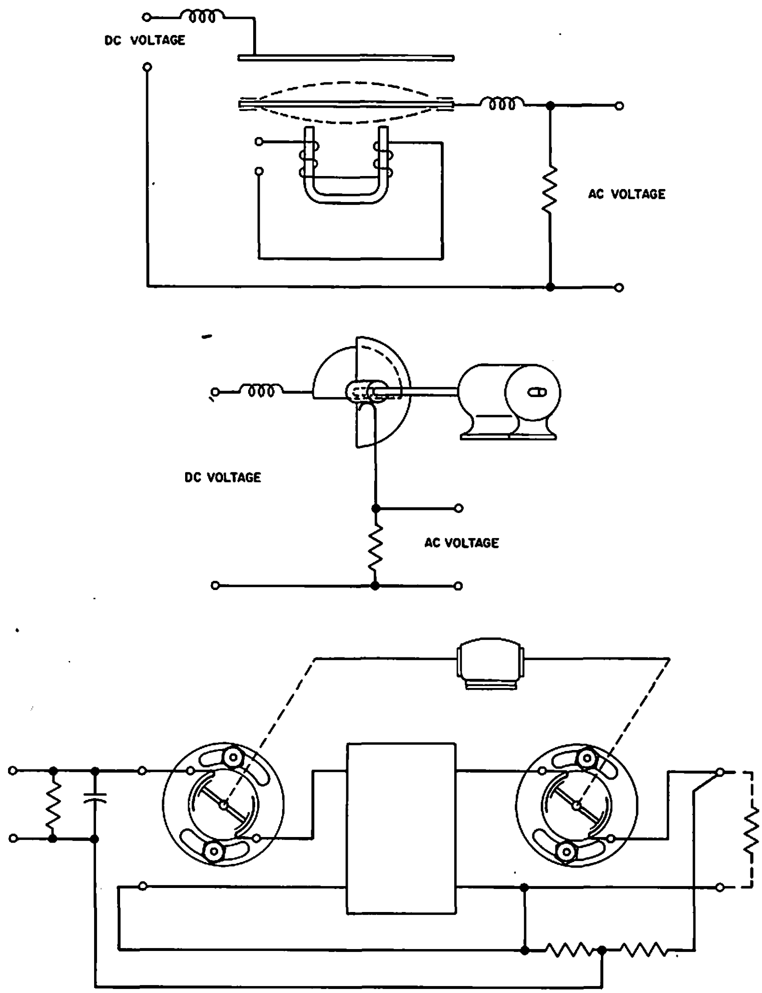


FIGURE 2. *Eberhardt contact modulation devices.*

illustrations appear in Figure 2. Other early patents were filed by F. S. Moseley, who describes chopper modulation and demodulation, and by Peterson, who showed a chopper servo system. But, see the tabulation appearing later in this chapter.

[2] Device for Amplifying Direct Voltage or Current, R. Eberhardt, et al, U.S. Patent No. 2,297,543, filed August 3, 1939, presently assigned to the Alien Property Custodian.

The rotary switch as a chopper, appears as noted in 1918, and its use is often still contemplated. It is certainly a usable method and is sometimes employed. However, it combines all the difficulty of a motor with all of those of a chopper. Usually the size and weight exceed that of a chopper. The sliding contact tends to generate quite a bit of noise, and unless the motor is synchronous, a phase relationship to the power frequency cannot be maintained.

FIGURE 3. *Early Airpax chopper Type A-589.*

One of the earliest choppers to be used in large volume was the Airpax Type A589, shown in Figure 3. This (one of the first 400 cycle choppers to exhibit a good order of reliability) is still in use. It, and its companions, Types A580 and A520, were used in 1948 and 1949 in jet fuel control and radar fire controls. Thus, they saw early service in the Korean war and took an important part in the performance and fire power of the fighter aircraft.

FIGURE 4. *Airpax high temperature (200° C) chopper Type 100.*

FIGURE 5. *VHF coaxial chopper Type 199.*

FIGURE 6. *Airpax Type 370 center pivoted chopper operates under high vibration and shock.*

Development proceeded in a number of directions as the art became more demanding. A demand for jet engine control led to the high temperature type, shown in Figure 4, Type 100, capable of operation up to 200° C, to Airpax Type 199, a coaxial chopper used at VHF, Figure 5. Small size will always be in demand, of course, as in Airpax Type 300. But, the chopper must operate under vibration and shock, answered by Type 370, Figure 6. It is safe to assume that chopper designs will get still smaller, more reliable and that performance will get better.

The role of a prophet is always risky. Nevertheless, certain facts are self-evident. The electromechanical chopper is not likely to disappear. As amply pointed out, it is uniquely capable of multiplication by almost zero and almost 1. Transistor choppers have invaded the art, but as they did the science moved on to greater precision. Thus while transistor devices can reach to 100 microvolts, electromechanical choppers move down to 1 microvolt. Since the reliability of moving contact devices has also become exceedingly good, it is reasonable to expect that electromechanical choppers will see many more years of use.

The literature of the chopper art is pretty extensive. Our following tabulation of articles includes information on how to avoid choppers, how to use choppers, a variety of chopper substitutes, and on allied subjects. It is reasonably complete in the English language. No such claim for completeness is made for the patents listed, nor, as has been pointed out earlier, are we always able to decide what is claimed to be invented. Patents and articles are both arranged in an approximate chronological order.

SECTION II

CHRONOLOGICAL TABULATION OF CHOPPER ARTICLES

THERMAL AGITATION OF ELECTRICITY IN CONDUCTORS, J. B. Johnson, Physical Review, Vol. 32, July, 1928, pp. 97-109.

RESISTANCE FLUCTUATIONS IN CARBON MICROPHONES AND OTHER GRANULAR RESISTANCES, C. J. Christensen and G. L. Pearson, Bell System Technical Journal, Vol. XV, pp. 197-223, April, 1936, also Monograph B922. Voltage fluctuations which occur in resistance elements of the granular type when a direct current is flowing are measured in the granular carbon microphone, commercial grid leaks and sputtered or evaporated metal films.

MAGNETIC SHIELDING OF TRANSFORMERS AT AUDIO FREQUENCIES, W. G. Gustafson, Bell System Tech. Journal, 17, p. 416, 1938.

SURVEY OF DC AMPLIFIERS, M. Artzt, Electronics, Vol. 18, p. 112, August, 1945.

A CONTACT MODULATED AMPLIFIER TO REPLACE SENSITIVE SUSPENSION GALVANOMETERS, M. Liston, et al, Review of Scientific Instruments, Vol. 17, No. 5, May, 1946. Describes a transformer type DC amplifier with chopper modulator and demodulator. Motor driven contacts operate at 75 cycles.

ELECTRONICALLY BALANCED RECORDER FOR FLIGHT TESTING AND SPECTROSCOPY, A. J. Williams, W. R. Clark and R. E. Tarpley, Transactions of the AIEE, 1946, Vol. 65, p. 416.

DC AMPLIFIER STABILIZED FOR ZERO AND GAIN, Williams, Jr., Tarpley and Clark, AIEE Transactions, p. 47, 1948, Vol. 67. Describes chopper amplifier using input modulator, both transformer and grid inputs for low and high impedances, uses chopper demodulation. Discusses output filter, feedback, electronic servo connection, problems with thermal junctions, input transformer design, circuit noise, oscillation, response time. One of the foundation articles for DC amplifiers.

STABILIZATION OF WIDE-BAND DIRECT-CURRENT AMPLIFIERS FOR ZERO AND GAIN, Goldberg, RCA Review, p. 296, June, 1950. A method for automatically stabilizing direct-current amplifiers against zero offset voltage and voltage drift. Appears to be the first publication on chopper stabilized amplifiers, sometimes described as Goldberg amplifiers. Uses a non-polarized vibrator driven at 193 cps, delivering 386 cycles. This operational amplifier appears in one of the early RCA computers, Typhoon.

DC AMPLIFIER USING AIR-COUPLED CHOPPER, Shafer, Electronics, March, 1950. A speaker driven, air coupled chopper is used to obtain low noise levels.

SUPERREGULATED POWER SUPPLIES, Vance & Shumard, Electronics, December, 1951. Uses chopper servos to stabilize amplifiers.

UNIVERSAL DIRECT-COUPLED DIFFERENTIAL AMPLIFIER, Goldberg, Electronics, October, 1951. Describes a differential first stage with both grids accessible, output reads the difference.

DC AMPLIFIER WITH REDUCED ZERO OFFSET, McAdam, Tarpley & Williams, Jr., Electronics, August, 1951. Chopper input and output amplifier with one microvolt offset at one megohm impedance. Discusses feedback and noise problems, investigates offset in input one megohm resistor. A rather complete discussion of noise and offset problems.

AUTOMATIC STABILIZATION OF HIGH-IMPEDANCE DC AMPLIFIER, Smith, Electronics, February, 1951. Brown DC amplifier and position servo, used to zero adjust a DC amplifier.

THE REQUIREMENTS AND DESIGN FOR A DIRECT-CURRENT NULL DETECTOR, Maltby, AIEE Technical Paper 51-344. Maltby describes the use of a DPDT chopper and presents an analysis of circuit performance.

RECENT DEVELOPMENTS IN MEASURING SYSTEMS FOR RECORDERS AND THEIR APPLICATION, Reprint from Transactions of Measurements Conference, Broomell and Emerich (1952). Describes several chopper circuits, also the L & N chopper.

CARRIER-COMPENSATION OF SERVOMECHANISMS USING SYNCHRONOUS VIBRATORS, M. I. T. Thesis, W. I. Cook, for Master's Degree. The purpose is to determine the feasibility of using choppers as components in the compensation of servomechanisms.

RATIO METER MEASURES REFLECTION COEFFICIENT, Rosenthal, Potter and Badoyannis, Electronics, November, 1952. AC ratio meter obtains DC performance by using chopper.

SUBMINIATURIZATION OF SERVO AMPLIFIER, Smith, Tele-Tech, November, 1952. Discusses the Minneapolis-Honeywell miniature chopper.

HIGH STABILITY COMPUTER REFERENCE POWER SUPPLY, Bradley and McCoy, Tele-Tech, June, 1952. Chopper amplifier stabilizes a regulating power supply used on REAC computer.

DRIFTLESS DC AMPLIFIERS, Bradley & McCoy, Electronics, April, 1952. Chopper stabilized amplifier used in Reeves REAC computer. Describes operational arrangements for summing, integration, etc.

REGULATED 1,600-AMPERE FILAMENT SUPPLY, Vance & Shumard, Electronics, February, 1952. Chopper amplifier amplifies DC error signal for thyratron control.

PARALLEL-TUNED CIRCUIT PERIODICALLY SWITCHED TO A DC SOURCE, RCA Review, 1952, 13, p. 386, L. J. Giacoletto.

EIN HOCHEMPFINDLICHER GLEICHSPANNUNGSVERSTARKET MIT HOHEM EINGANGWIDERSTAND, W. Kroebel, Zeitschrift fur Physik, 1952, 133, p. 30.

AN ISOLATING POTENTIAL COMPARATOR, T. M. Dauphinee, Canadian Journal of Physics, 31, p. 577, 1953.

ELECTROMECHANICAL MODULATOR FOR AN ANALOG COMPUTER, M. I. T. Thesis for Master's Degree, 1953, R. A. Gaskill. The development of an electromechanical capacitor modulator for use as a conversion unit in the M. I. T. flight simulator is described.

STEP-SWITCH CONVERTER, R. R. Bennett and H. Low, Electronics, November, 1953. Chopper amplifier used for accuracy. Stepping switches convert analog voltage to digital information.

A SURVEY OF THE LIMITS OF DC AMPLIFICATION, Verhagen, Proceedings of the I. R. E., p. 615, May, 1953. Dynamic and static balance of DC amplifier, compensation of tube differences, changes with time, shock stability, thermal effect, cathode changes.

DISPOSITIF ELECTRONIQUE POUR LA MESURE PRECISE DES TENSIONS CONTINUES PAR COMPARAISON, R. Aumont and Jacques Romand, Revue Génerale De L'Electricité, April, 1953. Chopper amplifier compares standard and unknown voltages to obtain precision.

A METHOD FOR THE AMPLIFICATION OF EXTREMELY SMALL THERMOELECTRIC VOLTAGES, W. Kroebel, Zeitschrift fur Angewandte Physik, 1953, 5, p. 286.

SERVOCOUPLER MATCHES AIRCRAFT ANTENNAS, Schwittek, Electronics, October, 1954. Phase sensitive RF discriminators feed two chopper servo amplifiers which may vary L and C of antenna network.

ANALYSIS OF SPECIAL PURPOSE FILTER INCORPORATING A PERIODICALLY CONDUCTING BILINEAR ELEMENT, Bolie, Proceedings of the I. R. E., September, 1954. Mathematical analysis of a filter network consisting of a resistor-capacitor-chopper combination. Amplitude modulated carrier is filtered, transmitted as a series of amplitude modulated rectangular pulses.

ZERO STABILIZATION OF DIRECTLY-COUPLED AMPLIFIERS, E. H. Frost-Smith and A. R. B. Churcher, The Elliott Journal, Vol. 2, No. 1, p. 136, August, 1954. A satisfactory method for insuring stability is the

use of a second amplifier in a feedback arrangement, the necessary characteristics of such an amplifier are here examined and some ways discussed of achieving them.

STABILIZED NOISE SOURCE FOR AIR-WEAPONS DESIGN, Beecher, Bennett, and Low, Electronics, July, 1954. Obtains Gaussian noise over 0-35 cps, to simulate air turbulence, etc., for air weapons design. A chopper rectifies output of 400 cycle band pass amplifier. Use of chopper avoids DC offset in output, likely to be present with other demodulators.

ALTITUDE CORRECTOR TO TRACKING RADARS, McQuiston, Electronics, June, 1954, a chopper position servo.

TIME-SHARED AMPLIFIER STABILIZES COMPUTERS, Slaughter, Electronics, April, 1954. A starved input DC amplifier stabilizes 30 DC operational amplifiers by a motor driven switch.

STABLE POWER SUPPLIES FOR MICROWAVE STANDARDS, Ernst, Electronics, January, 1954. Uses mercury cell reference and a chopper amplifier.

UNIVERSAL METER FOR MEASURING VOLTAGES AT HIGH IMPEDANCES, MICROMICROAMPERES AND INSULATION RESISTANCE, Clark, Watson & Mergner, Electrical Engineering, p. 41, January, 1954. Chopper type DC amplifier. Signal is fed back across indicating meter to an electronic null. Measures voltages up to 50 at 10^{10} ohms, current to 5×10^{-5} microamperes, and resistance to 10^{8} megohms.

HIGH SENSITIVITY DC BREAKER AMPLIFIER, Liston, Electronics, p. 206, January, 1954. A chopper amplifier with transformer input. "Breaker" frequency of 8 cps recommended. "Breaker" speeds over 80 cps reported undependable.

DC AMPLIFIERS, METHODS OF AMPLIFYING AND MEASURING SMALL DIRECT CURRENTS AND POTENTIALS, J. Yarwood and D. H. LeCroissette, Electronic Engineering, Vol. 26, p. 14, 1954. The principles of the single-valve amplifier for the amplification and measurement of small unidirectional currents and potentials are established.

VIBRATORY POWER CONVERTORS, R. H. Evans, Proceedings of I.E.E., Monograph No. 109R, September, 1954 (102 C, p. 62).

USE CHOPPERS FOR CONTROL, F. Rockett, Automatic Control, Nov., 1955. Describes Philbrick chopper amplifier and simple circuits.

TEMPERATURE INDICATOR FOR AIRCRAFT ENGINES, Koletsky, Electronics, p. 129, November, 1955. Chopper position servo, transformer input chromel alumel couple. A reference voltage is used.

DESIGN OF A RATIOMETER, Kuehn, Electronic Equipment, November, 1955. Describes Hycon 625 Ratiometer, chopper position servo drives indicator giving digital indication of voltage ratios.

THE HUSHED TRANSISTOR AMPLIFIER, W. K. Volkers, N. E. Pederson, a paper presented at the National Electronics Conference, October, 1955. Before the arrival of the "hushed" transistor amplifier, described in this paper, transistors had been considered to be inevitably more noisy than vacuum tubes.

INTEGRATING AMPLIFIERS BROADEN FIELD FOR APPLICATION OF ELECTRIC DRIVES, Beck, Electronics, October, 1955. Speed control uses integrating chopper amplifier to provide better speed accuracy, particularly at low speeds.

PORTABLE SECONDARY 1-MC FREQUENCY STANDARD, Kezer & Aronson, Instruments & Automation, October, 1955. Picks up circuits, etc., from P. G. Sulzer, High-Stability Bridge-Balancing Oscillator.

A STABLE DIODE CHOPPER CIRCUIT, H. Patton. A paper presented by the author at the Wescon convention in San Francisco, August, 1955. The stable diode chopper circuit operates at low millivolt signal levels; has a much greater life expectancy; and is small, compact and inexpensive. The circuit can be applied to phase detectors, time gates, suppressed carrier modulators, and transistor DC amplifiers.

LOW LEVEL THERMOCOUPLE AMPLIFIER & TEMPERATURE REGULATION SYSTEM, Dauphinne & Woods, The Review of Scientific Instruments, July, 1955. Mechanically driven 40 cps choppers modulate and demodulate transformer input amplifier.

HIGH STABILITY BRIDGE-BALANCING OSCILLATOR, Sulzer, Proceedings of the I. R. E., June, 1955. Series-resonance in a crystal exhibits high frequency stability. Error is introduced by amplifier (oscillator) phase shift. Crystal is used as resonant element in resonance bridge followed by a null amplifier controlling reactance tube. A chopper switches a capacitor from side to side of bridge, giving sense indication. A second chopper at output of 1 MC null detector delivers filtered DC to reactance tube for frequency correction.

CHOPPER-STABILIZED AMPLIFIER, Royce and Mathews, Electronic Equipment, May, 1955. Describes a KayLab chopper stabilized amplifier. Uses DPDT chopper.

TAPE CONTROLLED SERVOS SPEED CHEMICAL ANALYSIS, Donner, Electronics, February, 1955. Velocity servos using chopper amplifiers.

SPECIAL PURPOSE RELAYS GAIN NEW USES, Rockett, Electronics, February, 1955. Describes several types of special relays, includes choppers.

METHOD OF REDUCING ZERO ERROR AND DRIFT IN BREAKER TYPE DC AMPLIFIERS, T. M. Dauphinee, Review of Scientific Instruments, 1955, 26, p. 401.

VIBRATORY POWER CONVERTERS, Evans, R. H., Proceedings I.E.E. 102, Part C, March, 1955, pp. 62-80. (Monograph No. 109 R, September, 1954.) A full analysis of break-before-make buffered transformer-coupled circuits with particular reference to power supplies.

VIBRATORY POWER PACK TECHNIQUE, Sharp, L. W. D., British Communications & Electronics, October, 1955. Also reprint by Pressey Co. Deals with break-before-make SPDT transformer-coupled circuits with buffer capacitors, and shows waveforms during the free-oscillation period.

SWITCHING TRANSISTORS USED AS A SUBSTITUTE FOR MECHANICAL LOW-LEVEL CHOPPERS, Kruper, A. P., Communications & Electronics, March, 1955, No. 17, pp. 55-157.

TRANSISTOR CHOPPERS FOR STABLE DC AMPLIFIERS, Bright, R. L. & Kruper, A. P., Electronics, April, 1955.

MAGNETIC SHIELDING WITH MULTIPLE CYLINDRICAL SHELLS, W. G. Wadey, Review of Scientific Instruments, 27, 11, p. 910, November, 1956.

OPERATIONAL AMPLIFIER HAS CHOPPER STABILIZATION, D. Robinson, Electronics, pp. 182-185, September, 1956. An updated version is a part of the text of Part VII, The Contact Modulator.

DIGITAL PRESENTATION VACUUM-TUBE VOLTMETER, Nuut & Munsey, Electronics, January, 1956. Servo-driven self-balancing potentiometer with counter coupled to servo shaft provides a vacuum tube voltmeter with digital presentation. Chopper samples potentiometer signal and compares it with input.

A STUDY ON THE INPUT CIRCUIT OF THE CONTACT MODULATED AMPLIFIER, N. Kato & J. Ikenoue. Member Faculty Engineering, Kyoto University, 18 (2), April, 1956, pp. 112-129. In English — deals with transformer-coupled circuits whose primary receives DC pulses via an on-off switch.

TRANSISTORIZED LOW-LEVEL CHOPPER CIRCUITS, Richard B. Hurley, Electronic Industries & Tele-Tech, p. 42, December, 1956. Advantages of speed, ruggedness, and life expectancy more than compensate for lower switching "quality." Use of multiples and external "stabilizing" and "compensating" resistance is described.

ELECTRONIC DIFFERENTIAL ANALYZER OF THE G. M. DRZHYHANOVSKII POWER INSTITUTE OF THE U.S.S.R., I. S. Bruk, N. N. Lenov (Moscow) *Avtomatika i Telemekhanika,* Vol. XVII, No. 3, 1956 (In Russian). The analog computer equipment at this laboratory is described in some detail. Emphasis is placed on the operational amplifiers, multipliers and function generators. The operational amplifiers are stabilized with a 50 cycle chopper and by aging tubes and selecting for low

grid current, null drifts of .5uv/sec are obtained. The multiplier uses a 2.5 kc carrier sawtooth whose zero crossing phase can be varied by one input. This point in phase gates a pulse, amplitude modulated by a second input, thereby obtaining an average output proportional to their product. The function generator creates through harmonic mixing the desired periodic function in time. By sampling this wave at a phase linear with some input variable, the average output is the desired function of the input. Complete schematics and operating performance are given.

VIBRATOR AMPLIFIERS AND THEIR USES IN SERVOMECHANISMS, Ia. E. Gukailo, S. M. Fedorov, (Leningrad), *Avtomatika i Telemekhanika,* Vol. XVII, No. 10, 1956 (In Russian). Consider a slave driven vibrator which converts its drive coil AC power into a much higher level. By passing DC through a second drive coil on the vibrator, the unbalanced oscillations will produce a DC component in the load. This then offers a scheme for DC amplification at high power levels. Several SPDT and DPDT circuits are analyzed and their transfer functions derived. By using some feedback in the form of a self-driven coil, entrainment, or locked oscillator behavior is achieved which simulates backlash in the transfer function.

SERVO MODULATORS I, II, III and IV, Barber and Klivans, October-December, 1957, Control Engineering. This is an extensive omnibus type of article describing many types of modulation devices, including choppers. Has a tabulation of types and a lengthy bibliography.

A TRANSISTOR HIGH-GAIN CHOPPER TYPE DC AMPLIFIER, G. B. Chaplin, A. R. Owens, The Institution of Electrical Engineers, Paper No. 2442M, November, 1957. The paper describes a modulating system consisting of a transistor input chopper, a high-gain transistor AC amplifier and a transistor output chopper.

THE USE OF CHOPPERS IN DC AMPLIFICATION, P. T. McCauley, Research & Engineering, October, 1957.

A TRANSISTOR DC CHOPPER AMPLIFIER, Burton, Electronic Engineering, August, 1957. Transistor chopper multivibrator driven at 1.2 kc modulates thermocouple signal to transistor amplifier. Claims 1 mv stability from 12° to 50°.

SOME TRANSISTOR INPUT STAGES FOR HIGH-GAIN DC AMPLIFIERS, G. B. Chaplin, Institution of Electrical Engineers, Paper No. 2382M, July, 1957. The sensitivity of a DC amplifier using transistors is mainly limited by the variation with temperature of the inter-electrode potentials and currents of the transistors, the variation between any two temperatures being termed the thermal drift. The use of the transistor in a common-emitter chopping circuit results in a current drift of 50 microamperes and a voltage drift of only 2 millivolts, while reversing the functions of emitter and collector reduces these drifts to 3 micro-

amperes and 100 microvolts, respectively. The latter voltage drift is considerably less than that of most thermionic-valve amplifiers, but the current drift is still large and limits the use of the chopper to low-impedance sources.

A STUDY OF THE TRANSFER FUNCTION OF CONTACT-MODULATED AMPLIFIERS, Krantz, Salati & Berkowitz, Transactions of AIEE, March, 1957. Contact modulators are particularly suitable for several types of DC and AC amplifiers, and DC servomechanisms. Diagrams basic amplifier, modulator, AC amplifier, demodulator, filter, feedback. Derives expressions for the transfer function for the average value of modulated chopper wave, and for combination, also for half and full wave modulation.

PROPERTIES OF SOME DC-AC CHOPPER CIRCUITS, I. C. Hutcheon, Institution Monograph No. 218-R, January, 1957. The paper analyses five basic types of chopper circuits suitable for converting low-level direct voltages to alternating voltages. Three of them are transformer coupled; all have resistive loads.

SILICON DIODE CHOPPER STABILIZED DC AMPLIFIER, Fleming, Electronics, January, 1957. Claims impedance of about one megohm and stability of 100 microvolts per hour (at room ambient). Uses tube demodulator.

ELIMINATION OF AC BEATS IN AN ISOLATING POTENTIAL COMPARATOR CIRCUIT, Dauphinee, T. M., The Review of Scientific Instruments, Vol. 28, No. 6, June, 1957, p. 467. Choke filter and three transfer capacitors at 120° relative phasing.

DRIFT-CORRECTED DC AMPLIFIER, M. H. McFadden, Electronic and Radio Engineer, October, 1957, Vol. 34, No. 10, pp. 358-364.

LOW LEVEL TRANSISTORIZED CHOPPER AMPLIFIER, H. F. Harris and T. E. Smith, IRE Trans Telemetry and Remote Control, Vol. TRC-3, No. 1, paper 3-4, 8 pp. (April, 1957). The chopper is capable of processing 900 signals per second at a common mode rejection of 1000:1. The gain is smoothly adjustable ± 20% of full scale and the output is linear within 1% of a straight line drawn between the end points. Operation is consistent over temperature range of + 40°F to 165°F with less than 10% variation in gain. The rise time is less than 300 microseconds. Input impedance is 25 k ohms or better and output impedance is 5 k ohms or less. Ripple is less than 10 millivolt peak.

A SENSITIVE SUPERCONDUCTING "CHOPPER" AMPLIFIER, A. R. DeVroomen and C. Van Baarle, Physica XXIII (The Physica Foundation, Amsterdam), pp. 785-794, 1957. A superconducting "chopper" amplifier without movable parts and possessing a direct current sensitivity better than 10^{-11} volt is described. Compactness of cryostat equipment

and convenience of observation were factors controlling design. The instrument was developed in particular for the study of thermoelectricity in metals at low temperatures.

CONSIDERATIONS OF RELAY DYNAMICS WITH AN EXAMPLE OF NON-LINEAR VIBRATING REED DESIGN, David A. Robinson, presented at Fifth National Conference on Electro-Magnetic Relays, Oklahoma Inst. of Tech., Oklahoma A & M College, Stillwater, Oklahoma.

A LOW LEVEL, HIGH ACCURACY DC MAGNETIC AMPLIFIER, B. Mazzeo, Electrical Manufacturing, November, 1958. Magnetic amplifiers can be used to amplify thermocouple and other DC transducer voltages with high accuracy and reliability. Performance similar to chopper amplifiers is obtained. Complete isolation is possible between transducer and load.

DESIGN AND APPLICATION OF A SYNCHRONOUS CONVERTER, Part I and Part II, British Communications and Electronics, I. E. Hutcheon, Part I, p. 512, Vol. 5, July, 1958; Part II, p. 602, August, 1958. Part I, design of the synchronous converter of George Kent, Ltd. is a plug-in resonant-reed chopper intended for industrial use. Basically a SPDT switch for use with circuit voltages from zero to 10 volts. Part II discusses design of equipment associated with the chopper and circuit applications.

USE OF NOISE AND STATISTICAL TECHNIQUES IN ANALOG COMPUTATION, Chapter 26 in HANDBOOK OF AUTOMATION, COMPUTATION AND CONTROL, Vol. 2, H. Low, Computers and Data Processing, to be published by John Wiley & Sons. Circuits described use chopper as demodulator.

A STUDY OF THE ANALYTICAL REPRESENTATION OF A SYSTEM WITH FEEDBACK AND MODULATION, W. E. Nesbitt (D.E.), Doctor Thesis, 1958, Johns Hopkins University. Doctor thesis analyses feedback systems involving carrier, has chopper stabilized DC amplifiers.

PERFORMANCE CALCULATIONS FOR DC CHOPPER AMPLIFIERS, I. E. Hutcheon, Electronic Engineering, August, 1958. It is shown how input resistance, output resistance and overall DC gain can be calculated for a chopper amplifier employing any arrangement of perfect switches, resistors and capacitors in the chopping and rectifying circuits.

THE TRANSISTOR REGULATED POWER SUPPLY AS A FEEDBACK CONTROL SYSTEM, Norton & Jamison, Jr., Electromechanical Design, June, 1958. Brief mention of chopper modulation for best precision.

DIGITAL VOLTMETERS, (Editor), Electromechanical Design, June, 1958. Editorial describing digital voltmeter made by Non-Linear Systems, Inc. Uses chopper amplifiers.

TRANSISTOR CHOPPER, R. H. Williams, Electronics, May, 1958. Transistorized control circuit including a frequency-determining crystal oscillator feeds a voltage chopper which doubles the 28 volt DC supply and divides

a driving frequency. The AC pulsed output then drives a synchronous clock motor. System measures time intervals from 10 seconds to minutes with an accuracy of 0.02% over a temperature range of -55 to $+70°C$.

A STABLE GAIN TRANSFORMER COUPLED TRANSISTOR A. F. AMPLIFIER, H. Kemhadjian, Mullard Technical Communications, January, 1958. This article describes the design of a transistor DC "chopper" type amplifier using cascaded transformer-coupled grounded base transistor stages instead of the usual grounded emitter stages.

NEW TYPES OF DC AMPLIFIER, Part I, THE CASCADE-BALANCE SYSTEM, D. R. Martin, Electronic & Radio Engineers, January, 1958. THE REFLEX-MONITOR SYSTEM, PART II, D. R. Martin, Electronic methods of drift correction in a DC amplifier. A periodic drift correction using relays, takes amplifier out of use while connection is made. Choppers could be used; 5 cps is suggested. An extension of the Owen-Prinz method uses 2 identical amplifiers. Part II extends this to a balanced input.

SOME ASPECTS OF CONTINUOUS pH MEASUREMENT, Brooks, D., Medlock, R. S., Rudd, D. A., Instrument Engineer, Vol. 2, No. 6, October, 1958. (Also G. Kent, Ltd., Publication TP. 5036). Mainly chemical, but deals briefly with electrometer unit using transfer capacitor for DC isolation.

CONTROL SYSTEM COMPONENTS, Chapter 2, DC Amplifiers, Gibson & Tuteur, McGraw-Hill Book Company. The authors discuss DC amplifiers with and without modulator systems.

DC TO AC MODULATORS, George Sideris, Electronics, p. 47, January 23, 1959. Table gives relative performance that can be expected from nine different types of modulation devices.

DIFFERENTIAL CHOPPER AMPLIFIER HAS HIGH INPUT IMPEDANCE, Dr. Franklin F. Offner, Electrical Design News, March, 1959.

GUIDANCE SYSTEMS IN MANNED SPACE FLIGHT, Cap and White, Electronics, p. 49, August 14, 1959. Inertial reference system uses a chopper amplifier with a gain of 250,000.

AMPLIFIERS FOR STRAIN GAGES AND THERMOCOUPLES, Burwen, Electronics, p. 43, July 24, 1959. Describes a chopper stabilized differential wide band amplifier.

A HIGH-SPEED, AIRBORNE DIGITAL DATA ACQUISITION SYSTEM, Cogan & Hodder, p. 117, September, 1959, IRE TRANSACTIONS, Professional Group on Space Electronics and Telemetry. System uses chopper at 1700 cps. Article discusses practical application to data systems.

FUNCTION GENERATION WITH OPERATIONAL AMPLIFIERS, Koenner and Konn, Electronics, p. 66, November 6, 1959.

SUBAUDIO TUNABLE AMPLIFIER, Reece, Electronics, p. 72, November 6, 1959.

SECTION III

PATENTS

CHOPPER CIRCUITS

No. 1,378,712, filed January 17, 1918, J. W. Milnor, AMPLIFIER. Rotary commutator devices are used for modulation and demodulation, permitting amplifier use on telegraph signals of 10-20 cycles. Audions and transformers are used. Patented 1921 and assigned to Western Union. Object was to amplifiy weak telegraph signals on submarine cables. A tuned vibrating reed is used as rectifier in a variation. Also a dry rectifier. Output drives telegraph sounder. Filter used also.

No. 2,133,670, filed November 8, 1935, H. Schuchmann. VOLTAGE MEASURING SYSTEM. Echo depth sounder application. Time interval measured by amount capacitor discharges in that time. Charge on capacitor compared to standard by means of chopper.

No. 2,114,298, filed November 19, 1935, R. Gunn. APPARATUS FOR INDICATING SMALL ELECTROMOTIVE FORCES. Motor driven commutators modulate and demodulate the input and output of an amplifier.

No. 2,225,700, filed December 20, 1937, G. F. Laing, THERMOCOUPLE APPARATUS. Chopper compares measuring thermocouple to reference thermocouple in oil burner control application.

No. 2,323,966, filed October 7, 1938, Maurice Artzt, AMPLIFIER. Assigned to RCA. Chopper modulates DC signal from photoelectric cell.

No. 2,297,543, filed August 3, 1939, R. Eberhardt, et al, DEVICE FOR AMPLIFYING DIRECT VOLTAGE OR CURRENT. Chopper modulation, DC feedback, rotary modulator, capacitor modulator, demodulator networks. Assigned to Alien Property Custodian.

No. 2,413,788, filed May 11, 1942, W. E. Sargeant, et al, AMPLIFIER FOR SMALL VOLTAGES. Assigned to General Motors. Cam driven chopper used to modulate thermocouple voltage and demodulate amplifier output.

No. 2,315,714, filed June 12, 1942, Harry S. Jones, MEASURING APPARATUS. Straightforward chopper application. Chopper modulates thermocouple voltage to be amplified and fed to indicating device.

No. 2,385,481, filed March 25, 1943, Walter P. Wills, MEASURING AND CONTROLLING APPARATUS. Assigned to Brown Instrument Co. Self balancing potentiometric measuring device. Chopper modulates signal from thermocouple bridge network, signal is amplified and used to drive the device to null.

No. 2,459,730, filed June 30, 1944, Albert J. Williams, Jr., MEASURING SYSTEM WITH CAPACITOR HAVING CHARACTERISTICS OF AN INFINITE CAPACITY. Assigned to Leeds & Northrup. Chopper is used to compare input from thermocouple to output voltage of device.

No. 2,700,135, filed August 25, 1944, W.E. Tolles, PRODUCT-TAKING SYSTEM. Apparently a basic patent on the ring modulator and ring demodulator. Assigned to the Government. (U.S. Navy.)

No. 2,459,177, filed March 5, 1945, F. L. Moseley, et al, DIRECT CURRENT AMPLIFIER. Chopper modulator and demodulator, transformer type. Assigned to U.S. Government. Patent looks pretty fundamental, however, note Milnor, 1918.

No. 2,510,930, filed October 11, 1945, Kenneth G. MacLeish, EXPANDED SCALE AMMETER. Assigned to U.S.A., as represented by AEC. Difference between voltage across 50 mv shunt and reference potential is amplified and fed to meter. The reference provides the expansion.

No. 2,624,778, filed October 25, 1945, G. J. Perlow, et al, ELECTRONIC FLUXMETER AND ALTERNATING CURRENT AMPLIFIER. Signal from underwater search device is chopper modulated. A mine detector. No assignment noted.

No. 2,574,656, filed May 14, 1946, J. B. Peterson, TRUE AIRSPEED INDICATOR. A chopper servo amplifier is used on the output of a potentiometer.

No. 2,456,420, filed August 31, 1946, E. J. Jackson, ELECTRONIC DIRECT CURRENT CONTROL SYSTEM. Assigned to General Electric. One set of contacts chops the DC input signal. The auxiliary contacts are connected in circuit between the cathodes of discriminator stage and one terminal of the discriminator supply. Chopper auxiliary contacts make in one direction only, giving uni-directional output.

No. 2,485,948, filed January 31, 1947, A. J. Williams, Jr., et al, LOW-FREQUENCY CONVERTER-AMPLIFIER SYSTEM. Assigned to Leeds & Northrup. Shows a variety of modulator circuits.

No. 2,497,129, filed May 2, 1947, Max D. Liston, RADIATION DETECTING APPARATUS. Assigned to Perkin Elmer Corp., Glenbrook, Conn. Chopper modulates signal from thermopile and demodulates amplifier output to give a meter display of radiation intensity.

No. 2,628,994, filed October 31, 1947, Louis Goodman Mills, SWITCHING ARRANGEMENT FOR ELECTRICAL MEASURING SYSTEMS. Assigned to Foxboro Co. Chopper used to modulate DC potential as well as several other functions in the recorder. Chopper used is a four pole unit.

No. 2,532,911, filed December 3, 1947, R. L. Henson, Jr., et al, DRIVE FOR SEISMOGRAPHIC RECORD STRIPS. Power vibrator is driven as

a chopper to drive a synchronous motor in order to obtain uniform chart speed.

No. 2,508,082, filed December 11, 1947, Sidney Wald, REMOTE CONTROL SYSTEM FOR DC MOTORS. Assigned to RCA. A DC potential variable in magnitude and polarity is added to or subtracted from reference potential. Difference signal is modulated by the chopper, amplified and used to control the motor.

No. 2,584,954, filed May 8, 1948, A. J. Williams, SELF BALANCING ELECTRICAL SYSTEM. Assigned to Leeds & Northrup. An AC recorder application wherein a chopper rectifies AC input for differentiation or integration and reconverts to AC. This provides a derived signal which is an infinitesimal calculus function of the original signal.

No. 2,622,231, filed September 28, 1948, John W. Gray, INTEGRATOR. Assigned to General Precision Laboratories. Choppers used to supply the two fields of a two phase motor. Circuit is arranged so that choppers can be biased to either side, causing a resultant torque in either direction.

No. 2,622,192, filed January 6, 1949, Raymond E. Tarpley, MEASURING SYSTEM WITH GRID CURRENT SUPPRESSOR. Assigned to Leeds & Northrup Co. Chopper compares input to position of recorder.

No. 2,540,825, filed January 22, 1949, J. M. Lafferty, DIRECT CURRENT AMPLIFICATION SYSTEM. A bridge modulator with a square wave drive. No choppers. Assigned to General Electric.

No. 2,656,498, filed March 19, 1949, P. S. Goodwin, ELECTROMETER NETWORK. A DC preamplifier feeds a chopper servo, driving a potentiometer controlling a recorder.

No. 2,684,999, filed April 28, 1949, E. A. Goldberg, et al, STABILIZED DIRECT CURRENT AMPLIFIER. Chopper stabilization of DC amplifier. Assigned to RCA.

No. 2,685,000, filed April 29, 1949, A. W. Vance, STABILIZED DIRECT CURRENT AMPLIFIER. Use of one chopper for both modulation and demodulation of Goldberg type of stabilized DC amplifier. Assigned to RCA.

No. 2,688,729, filed July 28, 1949, Franklin F. Offner, RECORDER AMPLIFIER. Chopper modulates and demodulates low frequency signal.

No. 2,711,500, filed October 12, 1949, F. L. Maltby, MEANS FOR ESTABLISHING THE PHASE OF AN ALTERNATING CONTROL VOLTAGE AND A FOLLOW-UP CONTROL SYSTEM EMBODYING THE SAME. Chopper performs straight-forward modulation and demodulation.

No. 2,724,022, filed February 7, 1950, A. J. Williams, et al, FAST ACTING FEEDBACK AMPLIFIERS FOR HIGH IMPEDANCE SOURCES. Assigned to Leeds & Northrup Co. A device for measuring voltage from

a high impedance source. Arrangements made to nullify the effect of instrument capacitance on the time response of the instrument.

No. 2,619,514, filed February 18, 1950, Leonard Stanton, CAPACITY REBALANCING TYPE MEASURING APPARATUS. Assigned Minneapolis Honeywell. One chopper switches C_1 (fixed capacitor) from thermocouple to measuring device. Other chopper connects variable capacitor C_2 to standard cell and measuring device. C_2 is adjusted until $E_1 = E_2$.

No. 2,618,674, filed February 18, 1950. See No. 2,619,514, filed February 18, 1950.

No. 2,648,037, filed February 18, 1950, Thomas R. Harrison, AN IMPROVED APPARATUS FOR MEASURING SMALL VARIATIONS IN THE RESISTANCE OF A VARIABLE RESISTOR. Assigned to Minneapolis Honeywell. Chopper compares voltage across unknown resistance to voltage across standard resistance.

No. 2,758,079, filed March 29, 1950, Edgar L. Eckfeldt, ELECTROLYTIC DETERMINATION OF THE CONCENTRATION OF A CONSTITUENT IN A FLUID. Chopper is used to modulate the DC signal from pH indicating electrodes.

No. 2,638,811, filed April 22, 1950, See No. 2,697,791, filed September 4, 1952.

No. 2,709,205, filed June 23, 1950, John Archibald Colls, DIRECT COUPLED THERMIONIC VALVE AMPLIFIER. Assigned to Southern Instruments. Limited, Fernhill, Hawley, Camberley, England. A high gain amplifier which will retain the DC component of a voltage to be studied and at the same time give good response at high frequency components. Chopper compares output to input to correct for drift. Second chopper operates as half wave rectifier on output.

No. 2,583,339, filed August 17, 1950, James C. Mouzon, SERVO MOTOR DAMPING SYSTEM. Assigned to Minneapolis Honeywell. Chopper used to modulate thermocouple voltage.

No. 2,593,950, filed December 6, 1950, A. J. Williams, MOTOR CONTROL SYSTEM. Assigned to Leeds & Northrup. Thermocouple potential compared to reference, difference is amplified and used to drive recorder motor.

No. 2,615,064, filed December 22, 1950, L. Stanton, AMPLIFIER FOR STABILIZING SMALL UNIDIRECTIONAL SIGNALS. Assigned to Minneapolis Honeywell. A device to provide a constant DC voltage from a commercial source of voltage which fluctuates; chopper is used to modulate the DC input.

No. 2,619,552, filed February 7, 1951, A. Q. Kerns, AUTOMATIC DRIFT CORRECTOR. Chopper amplifier corrects the offset of a DC amplifier. Assigned to U.S. Government.

No. 2,713,135, filed March 9, 1951, F. Sutherland Macklen, INTERPOLATION SERVO. Assigned to Servo Corporation of America. An improved servomechanism wherein position and rate of change of position outputs may be available from input signals representing periodic sampling of a given electrical signal. Sampling is done by means of chopper.

No. 2,648,015, filed May 31, 1951, Joseph F. Clayton, SYSTEM FOR MEASURING RADIOACTIVITY. Assigned to Bendix Aviation Corporation. Chopper modulates DC signal from radiation detection device to be amplified and rectified.

No. 2,707,255, filed June 27, 1951, Robert M. Byrne, CHOPPER MODULATED ELECTRIC MOTOR SERVO AMPLIFIER. Assigned to Goodyear Aircraft Corp., Akron, Ohio. An error detector and amplifier for a DC servo system in which the error detector is in the form of a chopper modulator which compares the input signal to the output signal of the servo amplifier.

No. 2,773,946, filed September 27, 1951, S. Greenberg, et al, DEVICE FOR DETECTING THE SENSE AND MAGNITUDE OF A DC SOURCE. Chopper modulator shorts input. Assigned I. T. & T., filed 1951.

No. 2,744,969, filed August 4, 1952, W. D. Peterson. Assigned to North American Aviation. A chopper stabilized amplifier capable of functioning with minimum drift of calibration or sensitivity for extended periods of time.

No. 2,697,191, filed August 5, 1952, William H. Wannamaker, Gerald C. Mayer, FOLLOW-UP TYPE OF MEASURING APPARATUS. Three pole chopper is used to connect one capacitor alternately across unknown voltage and input to amplifier. Second capacitor is connected to known voltage and amplifier on opposite half cycles. Known voltage is adjusted until equal to unknown. Assigned to Minneapolis Honeywell.

No. 2,697,791, filed September 4, 1952, Frederick H. Krantz, CONVERTER STABILIZING SYSTEM. Assigned to Leeds & Northrup Co. A system for operating "vibrators" so that the phase shift and dwell time are relatively independent of the amplitude of driving voltage.

No. 2,856,468, filed October 27, 1952, Clifford E. Berry, NEGATIVE FEEDBACK AMPLIFIER IN A MEASURING SYSTEM. Assigned to Consolidated Electrodynamics Corp. Simple chopper modulator application. Low level DC input modulated and amplified.

No. 2,764,867, filed December 6, 1952, Thomas P. Farkas, SAFETY CIRCUIT FOR THERMOCOUPLE UNIT OF JET FUEL CONTROL. Assigned to United Aircraft Corp. A very complex circuit employing four choppers performing various functions.

No. 2,743,374, filed April 29, 1953, H. S. McCreary, Jr., CIRCUIT STABILIZER. Assigned to U.S.A. as represented by the Secretary of

the Navy. A chopper is used to shunt a high impedance with a low impedance in order to decrease the objectionally long time constant and associated long transient such as in photo-multiplier circuits when supply voltage is changed.

No. 2,813,248, M. C. Ferre, filed May 21, 1953, ELECTRICAL WELL LOGGING. Assigned to Schlumberger Well Surveying Corp., Houston, Texas. Chopper used to switch input to instrument from electrodes in well to reference electrode at remote point. Well logging application.

No. 2,688,112, filed July 1, 1953, F. T. Wimberly, ELECTRICAL CIRCUITS. A vacuum tube modulator device for a DC amplifier.

No. 2,790,944, filed September 11, 1953, Abraham W. Siff, Kenneth R. Neale, SHIELDED MEASURING APPARATUS. Assigned to the Bristol Co. A device for measuring small DC potentials utilizing a chopper to modulate the input.

No. 2,728,858, filed October 26, 1953, G. F. Ziffer, REGULATED POWER SUPPLY. Assigned Tracerlab, Inc. Portion of output voltage is bucked out by standard cell. The error difference is chopper modulated, amplified and demodulated. Signal controls amplitude of the oscillator.

No. 2,795,653, filed November 12, 1953, Rawley D. McCoy, VACUUM TUBE VOLTMETER AMPLIFIER CIRCUIT. Assigned to Reeves Instrument Corporation. AC amplifier and DC amplifier connected in cascade. Chopper serves as modulator and demodulator as well as the means of inserting the feedback to obtain the required high input impedance.

No. 2,801,296, filed February 9, 1954, F. H. Blecher, DC SUMMING AMPLIFIER DRIFT CORRECTION. Drift compensation, using chopper and DC feedback. Assigned to Bell Telephone.

No. 2,826,733, filed March 16, 1954, Robert R. Steward, ELECTRICAL APPARATUS. Assigned to Minneapolis Honeywell. Chopper modulates thermocouple potential to be measured.

No. 2,846,523, filed October 29, 1954, Minard A. Leavitt, Ivan C. Lutz, AMPLIFIERS FOR AC SIGNALS WITH SUPERIMPOSED DC. Assigned to U.S.A. as represented by AEC.

No. 2,820,855, filed July 7, 1955, Solomon Sherr, HIGH IMPEDANCE TRANSISTOR AMPLIFIER. Assigned to General Precision Lab., Inc. Chopper compares voltage across resistor in series with base of transistor to a reference. The difference is amplified and used to control oven temperature in which transistor is located. The temperature is adjusted to the necessary value to maintain a constant base current.

No. 2,857,562, filed June 18, 1956, Ernst Umrath, A MECHANISM TO PROVIDE SIMPLE DAMPING ARRANGEMENT FOR THE MOVABLE MECHANISM OF SLIDE WIRE POTENTIOMETERS. Assigned

to Daystrom, Inc., Murray Hill, New Jersey. Chopper compares input to reference and difference is amplified.

No. 2,832,848, filed January 16, 1957, Glyn A. Neff, ELECTRICAL SIGNAL AMPLIFIERS. Chopper modulates low level DC input.

No. 2,471,252, issued May 24, 1949, Pierre Marie Gabriel Toulan, SINGLE STAGE HIGH GAIN AMPLIFIER. A rotary chopper is used as a modulator and demodulator in a DC amplifier.

No. 2,866,019, issued December 23, 1958, Norman E. Perersin, DRIFT-LESS DC AMPLIFIER. Chopper used is a 10 cps cam operated device. Invention provides a DC amplifier which will enable the use of a single power source without need for close supervision and control.

PATENTS
(BRITISH)

No. 657,690, British Patent, filed May 19, 1949, P. F. Blackman. Claims read directly on transformer coupled chopper amplifiers. Assigned to Sun-Vic Controls, Ltd., of London.

No. 664,279, British Patent, filed March 23, 1950, D. H. Parnum. Uses choppers.

AIRPAX ELECTRONICS has been prominent in the chopper field since 1948. Since that time Airpax has made innumerable contributions to the design, use and circuitry of electromechanical choppers. As new problems are encountered in modulator practice, the full weight and experience of the Airpax Engineering Staff is called upon to create a unit suitable for the specific application. There is never a question as to quality — Airpax traditionally manufactures products which set the standards for industry.

Airpax also produces transistor choppers, the more important types being listed in this catalogue. While transistor choppers find uses in certain wide-band applications, the inherent low noise, excellent null stability, and switch-like impedance change characteristics of the electromechanical chopper demand its use in applications where these considerations are important.

Chopper drivers, both vacuum tube and transistorized types, together with transformers used in chopper circuitry, round out the line of Airpax chopper devices. These items are also listed in the following catalogue pages.

Manufacturing procedures are rigidly detailed in a manner which permits no deviation. Quality Control exercises watchdog vigilance over every phase of production from incoming raw material inspection through packaging. Airpax customers are assured of choppers which are the ultimate in performance, reliability, and long trouble-free life.

Over 60 standard types are listed in the following catalogue pages and more than 275 types are now being, or have been, manufactured to customer requirements or specifications. Whatever your chopper requirement, Airpax can help you.

AIRPAX ELECTRONICS
INCORPORATED
CAMBRIDGE, MARYLAND
FORT LAUDERDALE, FLA.

CATALOGUE SECTION

SERIES 175 (60 CPS STANDARD TYPES)

These hermetically sealed, miniature choppers employ a 6.3 volt RMS, 60 CPS drive. Long life operation through extremes of environmental conditions is enhanced by proven low noise and chatter free performance. Basic design and structure are such that the SPDT switching action runs continuously in synchronism with the driving voltage. Applications include use in high speed servos, computers, DC bridges, time sharing devices, and in the comparison of two signals. They also serve as stabilizing elements in DC amplifiers and perform modulator-demodulator functions in auto-pilots and machine controls.

60 CYCLE STANDARD

6.3 VOLT, 60 CPS DRIVE . . . SPDT BBM CONTACTS

TYPE	SPEC.	MTG.	TEMPERATURE	DWELL	PHASE	NOISE
171	232-3	A	−40°+ 71°C	145° Min.	20°±5°	50uvAV
172	232-4	A	−40°+ 71°C	145° Min.	20°±5°	50uvAV
175	232	A	−65°+100°C	164°±14°	20°±5°	50uvAV
176	232-9	J	−65°+100°C	164°±14°	20°±5°	50uvAV
178	232-11	C	−65°+100°C	164°±14°	20°±5°	50uvAV
179	232-12	B	−65°+100°C	164°±14°	20°±5°	50uvAV

In the above chart and those following, "TYPE" is the Airpax part number. "SPEC" refers to the detailed Airpax performance specification against which these are built and tested. "NOISE" is given in terms of the specification limit, in microvolts or millivolts, and in terms of the method of measurement, as AV (average), PP (peak-to-peak). Vibration and temperature are the specification limits. "MTG" styles are shown on page 46. "MM" noted after mounting styles indicates mu-metal case. Unless noted otherwise all electromechanical choppers listed operate satisfactorily under vibration conditions over a range of 10 to 55 CPS.

RATINGS FOR SERIES 175 (60 CPS STANDARD TYPES)

ELECTRICAL CHARACTERISTICS

CONTACT RATING: Contacts are suitable for interrupting dry circuit signals and can handle surges as large as 2 milliamperes and 100 volts.

DRIVE: 6.3 volts RMS, at 60 CPS. Drive voltage may vary from 5.7 to 6.9 volts RMS. Drive frequency may vary from 57 to 63 CPS.

DWELL TIME: The moving contact is closed to a fixed contact for a minimum of 150 and a maximum of 178 electrical degrees except for Types 171 and 172 which have a 145° minimum. Dwell times on the two fixed contacts do not differ by more than 15 electrical degrees.

CHATTER: If present, chatter is confined to a very brief interval immediately after make and to another brief interval immediately before break and is no more than 4 degrees during each such interval.

PHASE ANGLE: 20 ± 5 electrical degrees at 25°C.

INSULATION: 100 megohms minimum (under standard atmospheric conditions) between contacts and ground and 10 megohms minimum from drive coil to ground.

NOISE: The induced or stray noise appearing between each contact and ground does not exceed 50 microvolts average across 1 megohm resistance with the chopper operating.

ENVIRONMENTAL CONDITIONS

Materials and methods of fabrication used in this Series of choppers assure their reliable operation under natural combinations of environmental conditions encountered in industrial and military service.

TEMPERATURE: −65°C to +100°C except for Types 171 and 172 which operate from −40°C to +71°C.

HUMIDITY: Hermetically sealed for operation at 100% RH up to +50°C with condensation as water or frost; seal prevents entry of contaminents such as fumes or moisture assuring trouble-free operation under any climatic condition.

VIBRATION: Operates under conditions of vibration of 0.060 inch total travel from 10 to 55 CPS; vibrations of 5 G up to 500 CPS have no effect on subsequent operation.

SHOCK: 100 G shocks in both directions along each principal axis for a total of 30 shocks produce no permanent damage.

ALTITUDE: Unaffected by atmospheric pressure up to 75,000 ft.

MECHANICAL CHARACTERISTICS

All choppers in the 175 Series are small, light-weight units, yet have contact rating comparable to those of larger units. The basic mechanism is available in a plug-in type or in a variety of solder-lug types.

LIFE: These choppers are capable of a minimum of 5,000 hours continuous operation. The circuitry used has a considerable bearing on performance, and it is thus difficult to specify life precisely.

SIZE: Approximately ¾ inch in diameter and a seated height of 1 7/16 inches.

WEIGHT: All choppers in the 175 Series weigh 1 oz. max. except Type 178 which weighs 1.5 oz.

HEADER: 7-pin miniature, compression glass bead insulators.

CASE: Deep drawn brass spun onto header and soldered to make a true hermetic seal.

FINISH: Gray enamel over zinc chromate primer. Withstands 50-hour salt spray test per MIL-STD-202A, Method 101A.

SERIES 300 (400 CPS STANDARD TYPES)

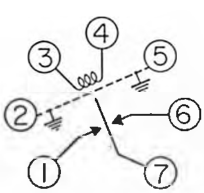
Airpax miniature Series 300 choppers are widely used as modulators and demodulators in stabilized DC amplifiers for analog computers and in servo-mechanisms for automatic controls — both industrial and military. Series 300 choppers are single-pole double-throw break-before-make resonant-reed contact modulators designed for handling information (low-level) signals.

All Airpax choppers have rather large contacts, and considerable spacing when in motion, providing stability against thermal and aging changes. Airpax choppers will handle surges as high as 100 volts at currents up to 2 milliamperes, as well as continuous dry circuit operation. Five thousand hours of operation is a reasonable life expectancy for a properly used Airpax chopper; life tests of well over 10,000 hours have shown only minor deviations from specification limits.

400 CYCLE STANDARD

6.3 VOLT, 400 CPS DRIVE . . . SPDT BBM CONTACTS

TYPE	SPEC.	MTG.	TEMPERATURE	DWELL	PHASE	NOISE
300	219	A	−65°+100°C	147°±18°	65°±15°	200 uv AV
302	219-7	C	−65°+100°C	147°±18°	65°±15°	200 uv AV
303	219-12	B	−65°+100°C	147°±18°	65°±15°	200 uv AV
307	219-18	AMM	−65°+100°C	147°±18°	65°±15°	200 uv AV
308	219-19	H	−65°+100°C	147°±18°	65°±15°	200 uv AV
309	219-20	HMM	−65°+100°C	147°±18°	65°±15°	200 uv AV



The majority of Airpax choppers have an SPDT BBM (break-before-make) contact arrangement. The construction of the chopper is such that there is no neutral position of the moving arm when the chopper is not being driven, and the movable contact will stop at random on either fixed contact when the drive is removed.

RATINGS FOR SERIES 300 (400 CPS STANDARD TYPES)

ELECTRICAL CHARACTERISTICS

CONTACT RATING: Contacts are suitable for interrupting dry circuit signals and can handle surges as large as 2 milliamperes and 100 volts.

DRIVE: At 400 CPS the chopper drive coil requires approximately 30 MA over the operating temperature range. Drive voltage may vary from 5.7 to 6.9 RMS volts. Drive frequency may vary from 380 to 420 CPS.

DWELL TIME: Moving contact is closed to a fixed contact for a minimum of 129 and a maximum of 165 electrical degrees. Dwell times on the two fixed contacts do not differ by more than 15 electrical degrees.

PHASE ANGLE: 65$\pm$15 electrical degrees.

CHATTER: If present, chatter is confined to a brief interval immediately after make and to another brief interval immediately before break and is no more than 4 degrees during each such interval.

INSULATION: 100 megohms minimum (under standard atmospheric conditions) between contacts and ground and 10 megohms minimum from drive coil to ground.

NOISE: 1.5 millivolts peak-to-peak.

ENVIRONMENTAL CONDITIONS

Materials and methods of fabrication used in Series 300 choppers assure their reliable operation under natural combinations of environmental conditions encountered in industrial and military service.

TEMPERATURE: −65°C to +100°C.

HUMIDITY: Hermetically sealed for operation at 100% RH up to 50°C with condensation as water or frost; seal prevents entry of contaminents such as fumes or moisture, assures trouble-free operation under any climatic conditions.

VIBRATION: Operates under conditions of vibration of 0.060 inch total travel from 10 to 55 CPS; vibrations of 5 G up to 500 CPS have no effect on subsequent operation.

SHOCK: 100 G shocks, in both directions along each principal axis for a total of 30 shocks, produce no permanent damage.

ALTITUDE: Unaffected by atmospheric pressure up to 75,000 ft.

MECHANICAL CHARACTERISTICS

All choppers in the 300 Series are small, light-weight units, yet have contact rating comparable to those of larger units. The basic mechanism is available either in a plug-in type or in a variety of solder-lug types.

LIFE: These choppers are capable of a minimum of 5,000 hours of continuous operation. The circuitry used has a considerable bearing on performance, and it is thus difficult to specify life precisely.

SIZE: Approximately ¾ inch in diameter and a seated height of 1 7/16 inches.

WEIGHT: Type 302 weighs 1.5 oz. max.; others in the Series weigh 1 oz. max.

HEADER: 7-pin miniature, compression glass bead insulators.

CASE: Deep drawn brass spun onto header and soldered to make a true hermetic seal. (Annealed mu-metal case available on special order.)

FINISH: Gray enamel over zinc chromate primer. Withstands 50-hour salt spray test per MIL-STD-202A, Method 101A.

SERIES 370 (400 CPS CENTER PIVOTED TYPES)

In the Airpax Series 370 chopper, the effects of shock and vibration are minimized by the unique, dynamically balanced armature structure. The armature is supported on a pivot at its center and is accurately balanced about this pivot. As a result, external mechanical vibration imparts only translation to the armature — the same as it does to the fixed contacts.

The drive coil and polarized magnetic circuit couple to the balanced armature at one end, while the moving contact is mounted at the opposite end, well removed from the magnetic field to minimize stray pickup.

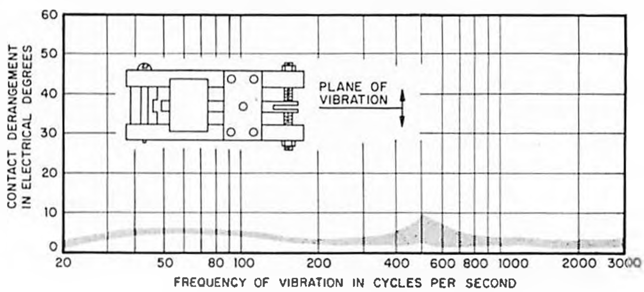
FOR HIGH FREQUENCY VIBRATION

6.3 VOLT, 400 CPS DRIVE . . . SPDT BBM CONTACTS

TYPE	SPEC.	MTG.	TEMPERATURE	VIBRATION	DWELL	PHASE
370	243	A	—65°+125°C	15G to 2500	160° ±15°	55° ±15°
371	243-1	I	—65°+125°C	15G to 2500	160° ±15°	55° ±15°
372	243-2	C	—65°+125°C	15G to 2500	160° ±15°	55° ±15°
373	243-3	B	—65°+125°C	15G to 2500	160° ±15°	55° ±15°

The noise appearing between any contact and ground is 200 microvolts or less.

EFFECT OF 15 G VIBRATION ON CONTACT ACTION



By supporting the moving contact at its center of gyration, Airpax engineers have produced a mechanical modulator that continues operating in the presence of vibration yet preserves the advantages of conventional choppers. These choppers provide 100% modulation, zero output with zero input, passive signal circuit, and wide dynamic signal range.

Contacts of any chopper are deranged from normal operation by shock and vibration. In Airpax Series 370 choppers, this derangement is below 10 electrical degrees for vibrations up to 15 G. The choppers are undamaged by vibrations up to 100 G.

RATINGS FOR SERIES 370 (400 CPS CENTER PIVOTED TYPES)

ELECTRICAL CHARACTERISTICS

CONTACT RATING: Contacts are suitable for interrupting dry circuit signals and can handle surges as large as 2 milliamperes and 100 volts.

DRIVE: At 400 CPS the chopper drive coil requires approximately 30 MA over the operating temperature range. Drive voltage may vary from 5.7 to 6.9 RMS volts. Drive frequency may vary from 380 to 420 CPS.

DWELL TIME: Dwell time is 145 degrees to 175 degrees duration, referred to a sinusoidal reference wave. Dwell time on one contact is balanced to within 15 degrees of the dwell time on the other contact.

PHASE ANGLE: 55 ± 15 electrical degrees at 25°C.

CHATTER: Chatter is limited to a brief period after make and to a brief period before break and does not exceed 4 degrees in any one period.

INSULATION: Insulation resistance as measured with a 200-volt source is at least 100 megohms between each contact terminal and case, and at least 10 megohms between each coil terminal and case.

NOISE: The induced or stray noise appearing between each contact and ground is less than 200 microvolts average across 1 megohm with the chopper operating.

ENVIRONMENTAL CONDITIONS

The Series 370 choppers use materials and methods of fabrication assuring reliable operation under any natural combination of environmental conditions encountered in industrial and military service.

TEMPERATURE: −65°C to +125°C.

HUMIDITY: Hermetically sealed for operation at 100% RH up to 50°C with condensation as water or frost; seal prevents entry of contaminents such as fumes or moisture, assures trouble-free operation under any climatic conditions.

VIBRATION: The chopper undergoes less than 10 electrical degrees of contact derangement when subjected to sinusoidal vibrations in any direction at an amplitude of 15 G over a frequency range of 10 to 2500 CPS. Such vibration produces no permanent change in the operation of the chopper.

SHOCK: The chopper is undamaged by 30 mechanical shocks of 100 G each applied 5 shocks in both directions of each principal plane.

ALTITUDE: Unaffected by atmospheric pressure up to 75,000 ft.

MECHANICAL CHARACTERISTICS

Series 370 choppers are small, lightweight units, yet have contact ratings comparable to those of larger units. The basic mechanism is available either in plug-in or a variety of solder-lug types.

SIZE: Approximately ¾ inch in diameter and a seated height of 1 7/16 inches.

WEIGHT: Type 372 weighs 1.5 oz. max.; Types 370, 371 and 373 weigh 1 oz. max.

HEADER: 7-pin miniature, compression glass bead insulators.

CASE: Deep drawn brass spun onto header and soldered to make a true hermetic seal.

FINISH: Gray enamel over zinc chromate primer. Withstands 50-hour salt spray test per MIL-STD-202A, Method 101A.

SERIES 2300 (400 CPS VERY LOW NOISE TYPES)

Series 2300 choppers, designed for 6.3 volt RMS, 400 CPS drive, achieve remarkable low noise levels, yet meet all electrical characteristics and environmental conditions of standard Airpax 400 CPS miniature choppers. Specific applications where these characteristics are particularly important include null seeking servos, instrumentation, instrument amplifiers, telemetering and control systems. Across a 100 ohm impedance, from any contact to ground, the noise level is four microvolts or less.

VERY LOW NOISE
6.3 VOLT, 400 CPS DRIVE . . . SPDT BBM CONTACTS

TYPE	SPEC.	MTG.	TEMPERATURE	DWELL	PHASE	NOISE
2300	249	J	−65°+100°C	150°±20°	30°±15°	4uv RMS
2302	249-7	CK	−65°+100°C	150°±20°	30°±15°	4uv RMS
2303	249-11	BK	−65°+100°C	150°±20°	30°±15°	4uv RMS
2307	249-16	K	−65°+100°C	150°±20°	30°±15°	4uv RMS

RATINGS FOR SERIES 2300

ELECTRICAL CHARACTERISTICS

DRIVE VOLTAGE AND FREQUENCY: Nominal 6.3 volts RMS ± 10% at 400 CPS ± 5%.

PHASE ANGLE: 30±15 degrees.

DWELL TIME: 130 to 170 degrees.

BALANCE: Dwell time on one fixed contact is balanced to within 18 degrees of dwell time on the alternate fixed contact.

NOISE: Four microvolts or less, RMS.

ENVIRONMENTAL CONDITIONS

TEMPERATURE RANGE: Continuous satisfactory operation from −65°C to +100°C.

HUMIDITY: Hermetic seal assures trouble-free operation under extremes of humidity and in any gaseous medium.

ALTITUDE: Unaffected by altitudes of 75,000 feet.

VIBRATION: 10 to 55 CPS.

SHOCK: Operation is satisfactory subsequent to 30 mechanical shocks of 30 G applied 5 shocks in both directions in each of 3 mutually perpendicular planes.

MECHANICAL CHARACTERISTICS

LIFE: Electrical characteristics remain unchanged for a minimum of 1,000 hours of operation.

SIZE: Diameter, ¾ inch; case seated height, 1½ inches; top terminals protrude approximately ¼ inch above case.

WEIGHT: 26 grams.

BASE HEADER: 7 pin miniature. Mounts in JAN TS102P01 socket with JAN TS102-U02 locking tube shield.

TOP HEADER: 2 terminal. Block centered on case top.

TYPES: Series 2300 choppers are also available with solder pin headers and solder lug top connectors and in mu-metal cans.

SERIES 310 (400 CPS HIGH TEMPERATURE TYPES)

Airpax Series 310 choppers are a high temperature version of the widely used Series 300 choppers. As such they are recommended for those applications requiring continuous reliable operation at temperatures from −65°C to +125°C.

Because of their switching action and low drive power, these high temperature choppers are operated with the same drive voltage (6.3 RMS volts) and frequency (400 CPS) as conventional types. Contacts are suitable for interrupting dry circuit signals and can handle surges as large as 2 ma. and 100 volts.

FOR 125°C AMBIENTS

6.3 VOLT, 400 CPS DRIVE . . . SPDT BBM CONTACTS

TYPE	SPEC.	MTG.	TEMPERATURE	DWELL	PHASE	NOISE
310	239	A	−65°+125°C	147° ±18°	65° ±15°	200uv AV
311	239-1	I	−65°+125°C	147° ±18°	65° ±15°	200uv AV
312	239-2	C	−65°+125°C	147° ±18°	65° ±15°	200uv AV
313	239-4	B	−65°+125°C	147° ±18°	65° ±15°	200uv AV
317	239-6	AMM	−65°+125°C	147° ±18°	65° ±15°	200uv AV
318	239-7	H	−65°+125°C	147° ±18°	65° ±15°	200uv AV

CHARACTERISTICS

Materials and constructional features are such that high or varying temperatures do not adversely affect the operation of these miniature choppers. The phase angle is only slightly affected by the electrical impedance of the drive coil and the mechanical phase of the resonant reed. Electrical phase changes with temperature due to the temperature coefficient of resistance of the copper drive coil wire. Mechanical phase changes with temperature due to the temperature coefficient of expansion of the metallic reed. The combined result is a slight decrease in overall chopper phase angle as operating temperature increases. The following table lists the ranges of phase angle of Series 310 choppers for the indicated temperature ranges.

Temperature in degrees C	Phase in electrical degrees
−65° to −20°C	67 ± 13
−20° to +100°C	63 ± 15
+100° to +125°C	57 ± 13

These spreads in phase angle are limits for all Series 310 choppers. Any one chopper can be expected to change in phase with temperature considerably less than indicated in the table.

SERIES 400 (400 CPS LEADS OUT THE TOP)

Drive coil leads exit through top of the enclosure in the 400 Series. This permits design flexibility and minimizes drive coil to contact coupling. Connections to associated components may be conveniently made above the chassis.

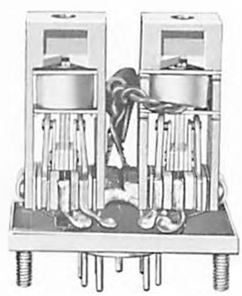
LEADS OUT THE TOP

6.3 VOLT, 400 CPS DRIVE . . . SPDT BBM CONTACTS

TYPE	SPEC.	MTG.	TEMPERATURE	DWELL	PHASE	NOISE
400	264	J	—65°+100°C	147°±18°	65°±15°	200 uv AV
402	264-4	CK	—65°+100°C	147°±18°	65°±15°	200 uv AV
403	264-5	BK	—65°+100°C	147°±18°	65°±15°	200 uv AV
404	264-6	K	—65°+100°C	147°±18°	65°±15°	200 uv AV

SERIES 600 and 800 (DPDT TYPES)

Airpax double-pole double-throw choppers, Type 600 for 400 CPS operation and Type 800 for 60 CPS operation, provide complete resistive and capacitive isolation. Performance is synchronous and common coil connections are used. Tracking between contacts and the balance of dwell times on the fixed contacts are closely controlled in the manufacturing process.



DOUBLE-POLE DOUBLE-THROW

6.3 VOLT DRIVE . . . MTG. NQ . . . DPDT BBM CONTACTS

TYPE	SPEC.	TEMPERATURE	DWELL	PHASE	NOISE	FREQ.
600	244	—65°+100°C	147°±18°	65°±15°	200uv AV	400
800	246	—65°+100°C	167°±20°	20°±8°	100uv AV	60

SERIES 360 (400 CPS LONG DWELL TIME)

Where circuit requirements call for a longer dwell time than obtainable with the standard Series 300, Airpax provides the Series 360. These choppers have a 10 percent greater dwell time and, like the Series 300, operate at 400 cycles per second.

400 CYCLE LONG DWELL TIME

6.3 VOLT, 400 CPS DRIVE . . . SPDT BBM CONTACTS

TYPE	SPEC.	MTG.	TEMPERATURE	DWELL	PHASE	NOISE
360	242	A	—65° +100°C	160° ± 15°	55° ± 15°	200uv AV
361	242-1	I	—65° +100°C	160° ± 15°	55° ± 15°	200uv AV
362	242-2	C	—65° +100°C	160° ± 15°	55° ± 15°	200uv AV
363	242-3	B	—65° +100°C	160° ± 15°	55° ± 15°	200uv AV
364	242-4	F	—65° +100°C	160° ± 15°	55° ± 15°	200uv AV
367	242-5	AMM	—65° +100°C	160° ± 15°	55° ± 15°	200uv AV

SERIES 190 (50 CPS STANDARD TYPES)

Choppers in the Series 190 are the 50 CPS equivalents of those in Series 175. Principally for overseas use, where fifty cycle power lines are common, they are used as modulators, demodulators, time sharing devices and in like applications.

50 CYCLE CHOPPERS

6.3 VOLT, 50 CPS DRIVE . . . SPDT BBM CONTACTS

TYPE	SPEC.	MTG.	TEMPERATURE	DWELL	PHASE	NOISE
190	234	A	—65° +100°C	167° ± 10°	20° ± 5°	50uv AV
192	234-2	C	—65° +100°C	167° ± 10°	20° ± 5°	50uv AV
193	234-3	B	—65° +100°C	167° ± 10°	20° ± 5°	50uv AV
194	234-4	F	—65° +100°C	167° ± 10°	20° ± 5°	50uv AV
196	234-5	J	—65° +100°C	167° ± 10°	20° ± 5°	50uv AV

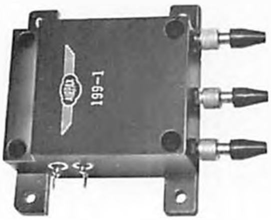
SERIES 1400 and 1800 (60 CPS MBB TYPES)

With a fairly low impedance source and high input resistance, considerable noise may appear during the chopper off time. Greater gain (with longer dwell time) is obtained by using Airpax MBB Types 1400 or 1800.

MAKE-BEFORE-BREAK

6.3 VOLT, 60 CPS DRIVE . . . NOISE IS 70 MICROVOLTS RMS OR LESS

TYPE	SPEC.	MTG.	TEMPERATURE	CONTACTS	DWELL	PHASE
1400	247	A	−40° +85°C	SPDT MBB	198° ± 20°	22° ± 10°
1800	252	NQ	−40° +85°C	DPDT MBB	200° ± 20°	22° ± 10°



SERIES 199 (100 CPS COAXIAL TYPES)

For use in the 200 to 400 MC range, the Type 199 chopper samples two signal sources for a single load or distributes a low level signal to two loads in a periodic manner. Switching frequency is 100 CPS.

COAX CHOPPERS FOR VHF

100 CPS DRIVE . . . DPDT BBM . . . VSWR 1.4 MAX.

TYPE	SPEC.	MTG.	DRIVE	TEMPERATURE	DWELL	PHASE
199-1	218-1	L	7.0 Sq	−55° +110°C	160° ± 15°	30° ± 10°
199-2	218-2	L	14.0 Sq	−55° +110°C	160° ± 15°	30° ± 10°

SERIES 5600 (400 CPS TRANSISTOR DRIVE TYPES)

These choppers, designed for transistor drive, have their drive coils tapped at the electrical center. No drive transformer is required which saves weight and space and improves the overall efficiency.

TRANSISTOR DRIVEN

9 VOLT, 400 CPS DRIVE . . . SPDT BBM CONTACTS

TYPE	SPEC.	MTG.	TEMPERATURE	DWELL	PHASE	NOISE
5600	286	A	−65° +100°C	147° ± 18°	65° ± 15°	1.5MV RMS
5606	286-1	Y	−65° +100°C	147° ± 18°	65° ± 15°	100uv RMS

REPLACEMENT TYPES

Replacement choppers in many shapes, sizes, and electrical characteristics are available. Stock is carried for replacement of many choppers of other manufacturers as well as those manufactured by Airpax.

These replacement types are designed to duplicate the function of the original type but incorporate many engineering improvements and more advanced manufacturing techniques.

These types are recommended for replacement service only. For new design requirements, select from the list of standard types to ensure the most advanced design and fastest delivery.

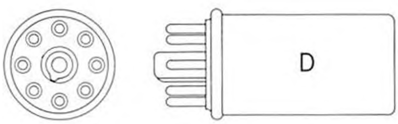
MUCH USED REPLACEMENTS

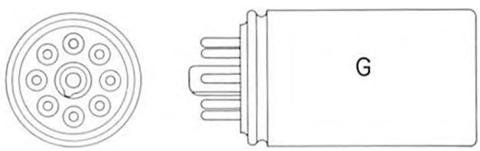
6.3 VOLT DRIVE . . . SPDT BBM CONTACTS

TYPES 170 AND 586R, 60 CPS   TYPE 589R, 400 CPS

TYPE	SPEC.	MTG.	TEMPERATURE	DWELL	PHASE	NOISE
170	232-1	D	−65° +100°C	165° ± 15°	20° ± 5°	100 uv AV
586R*	137	G	−65° + 85°C	158° ± 18°	40° ± 15°	2MV PP
589R*	104	G	−65° + 85°C	140° ± 25°	65° ± 15°	3MV PP

*Recommended for replacement only. (Suffix R indicates longer life, but otherwise same characteristics as older units of corresponding type number.)





AIRPAX TRANSISTOR CHOPPERS

Airpax has developed modulator-demodulator devices to a high degree of perfection through years of design and test experience. With the relatively recent demand for much higher switching frequencies, Airpax engineers have developed transistor choppers capable of linearly chopping or switching voltages over a wide dynamic range.

An inertialess device, the transistor chopper can be operated from drive voltages ranging from DC to as high as 100 KC for some models. The drive power required is very low, and can be either square or sine wave. At the higher chopping frequencies, drive power can be supplied by a transistor oscillator or its equivalent.

For guided missile and other military applications, Airpax transistor choppers are ideal. They are free from the effects of acceleration, shock, and vibration, and, because they are encapsulated, will operate at any altitude.

The transistor chopper provides reasonably good efficiency and linearity, and has a low null output. The noise figure is competitive with other types of modulator-demodulator devices in most applications and will not increase with aging.

The phase angle is approximately zero except at very high drive frequencies. With square wave drive, the dwell times are practically 180 degrees and, with symmetrical drive voltage, the dwell times are symmetrical.

Because of these superiorities, the transistor chopper has a distinct advantage in applications where phase shift must be held to a minimum, as in a quadrature rejection demodulator. Even with sine wave drive, the phase angle can be made very low by use of relatively high chopping voltage and a series resistor; however, care must be taken in such circuits to ensure that the peak inverse drive voltage does not exceed the chopper rating.

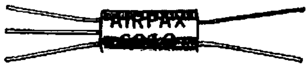
Airpax transistor choppers are available in SPST, SPDT, and full-wave modulator models. All transistors used are carefully selected and matched.

Airpax has accumulated extensive experience in transistor chopper circuitry. We suggest that you contact our Cambridge, Maryland Division.

TYPICAL APPLICATIONS:

- MODULATORS-DEMODULATORS
- DC AMPLIFIER STABILIZATION
- SERVO SYSTEMS
- DC MEASURING INSTRUMENTS
- OPERATIONAL AMPLIFIERS
- PORTABLE EQUIPMENT
- MULTIPLEXING EQUIPMENT
- TELEMETERING EQUIPMENT

SUBMINIATURE TRANSISTOR CHOPPERS



Reliability, miniaturization and low drive-power requirements are fully realized in this new line of choppers. Measuring only ½-inch in length, 3/16-inch in diameter, and weighing only one gram, Airpax subminiature transistor choppers fit the rapidly increasing demand for miniaturization and lightweight design advanced by the extensive use of transistors in the electronic industry. Suitable for most chopper applications, subminiature transistor choppers are particularly applicable in portable, telemetering, and other lightweight-design equipments where space is limited. Airpax subminiature transistor choppers are encapsulated to withstand shock and vibration. Because of their extremely small size, a separate drive transformer is required. Pigtail leads are provided for circuit connection.

Switching is performed without moving parts, permitting the transistor chopper to operate normally during shock and vibration. Chopper output is linear within 2% over a range of signals from approximately 5 millivolts to ⅓ the peak-to-peak drive voltage when square wave drive is used and from 10 millivolts to ⅛ the peak-to-peak drive voltage when sine wave drive is used with signal currents ranging up to 3 milliamperes.

SUBMINIATURE CHOPPER TYPES

DRIVE: 1 TO 20 VOLTS PK-PK, DC TO 100 KC, 20 MA MAX.
DWELL 175°±5° . . . PHASE 0°

TYPE	SPEC.	NOISE	MAX. VOLTS	SIGNAL SWITCHING	MA	TEMPERATURE
6010	291	150uvRMS	20V	SPST	3	−40°+85°C
			10V	SPDT		
6011	320	1MVRMS	20V	SPST	20	−40°+85°C
			10V	SPDT		

Types 6010 and 6011 have pigtail leads.

With the exception of Types 6010, 6011 and 6020-3 all transistor choppers are miniature plug-in types with self-contained drive transformers. Enclosure is a mu-metal case with a 7-pin header. Switching action for all types is BBM. Dwell times shown are for sine wave drive. Vibration: 20G to 2000 CPS for Types 6010, 6011 and 6020-3; all other types are 10G to 2000 CPS.

HIGH VOLTAGE TRANSISTOR CHOPPERS

Type 6025 is employed where high signal voltage levels are encountered. Type 6025 uses high voltage rated transistors and has a power requirement of less than 20 milliwatts. A linear output is achieved over a signal range from approximately 1 millivolt to 100 volts at signal currents to 10 MA.

HIGH VOLTAGE TYPES

DRIVE: 5.7 — 14 VOLTS RMS, 50-5000 CPS . . . DWELL 175°± 5°
PHASE 0°±5°

TYPE	SPEC.	NOISE	MAX. VOLTS	SIGNAL SWITCHING	MA	TEMPERATURE
6025	292	1MVRMS	100V	SPDT	10	−40° to +55°C

REPLACEMENTS
FOR ELECTROMECHANICAL CHOPPERS

Miniature Types 6015 and 6035 are for use in some equipments originally designed for electromechanical choppers. No circuit modifications are required and they may be plugged directly in as replacements. Input signal range of Type 6015 is from 1 millivolt to 5 volts at signal currents to 0.5 MA. Type 6035, a wider range version, handles input signals from the noise level to 20 volts at currents to 2.0 MA. Switching is BBM.

TYPE 300 REPLACEMENT

DRIVE: 5.7 — 14 VOLTS RMS, 50-5000 CPS . . . DWELL 175°± 5°
PHASE 0°±5°

TYPE	SPEC.	NOISE	MAX. VOLTS	SIGNAL SWITCHING	MA	TEMPERATURE
6015	295	3MVRMS	5V	SPDT	0.5	−40°+85°C
6035	351	1MVRMS	20V	SPDT	2	−40°+85°C
6036*	356	1MVRMS	20V	SPDT	1	−40°+85°C

*Type 6036 has a maximum null voltage of 150 uv.

HIGH TEMPERATURE TRANSISTOR CHOPPERS

Types 6020, 6021 and 6045 are high temperature silicon types and provide excellent characteristics to +125°C. These silicon transistor choppers are not as limited in temperature range and "off" resistance as are germanium types. SPST switching action is provided by Type 6020 which linearly modulates signals from 1 millivolt to 30 volts at signal currents to 3 MA. Types 6021 and 6045 are for shunt type switching circuits.

HIGH TEMPERATURE TYPES

DRIVE: 5.7 — 14 VOLTS RMS, 50-5000 CPS . . . DWELL 170°± 5°
PHASE 0°±5°

TYPE	SPEC.	NOISE	MAX. VOLTS	SIGNAL SWITCHING	MA	TEMPERATURE
6020	300	3MVRMS	30V	SPST	3	—55°+125°C
6021	355	1MVRMS	10V	SPST	1	—55°+125°C
6045	353	1MVRMS	15V	SPST	3	—55°+125°C

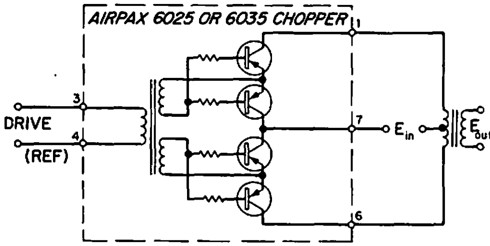
TRANSISTOR CHOPPER FOR PRINTED CIRCUITS

Type 6020-3 is a molded unit with four leads for printed circuit mounting and requires an isolation transformer, such as Airpax Type T-1817-1.

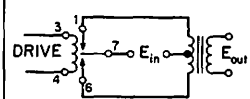
Type: 6020-3. Spec: 354; Drive: 2-20 V PK (20 MA Max.); Freq: DC to 100 KC; Dwell: 170°±5°; Phase: 0°±5°; Noise: 3 MV RMS; Max. Signal Voltage: 30 V; Signal Switching: SPST. 3 MA max; Temp: —55°+125°C.

TYPICAL CIRCUIT APPLICATIONS OF TRANSISTOR CHOPPERS

Any of the modulator circuits shown can be converted to demodulators by interchanging signal input and output terminals. Drive and signal voltage specifications are the same in either application.

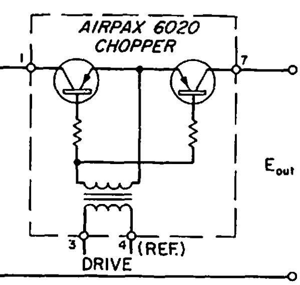


FULL WAVE MODULATOR

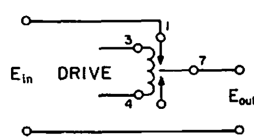


AIRPAX ELECTRO-MECHANICAL EQUIVALENT

FULL WAVE MODULATOR: This circuit is very efficient, supplying an output during the complete driving cycle, and as illustrated, a Type 6025 or Type 6035 is employed. The 6015 may also be used in this circuit with an attendant increase in the offset component. The chopper signal voltage may be as high as the voltage rating of the 6025 and 6035 transistor choppers. If Type 6015 is used, the maximum signal voltage must not exceed the peak of the driving voltage, and for good linearity, should be much less. Types 6015, 6025, and 6035 provide a differential input to the associated amplifier. Frequency and impedance values will depend to a considerable extent on the amplifier input transformer.

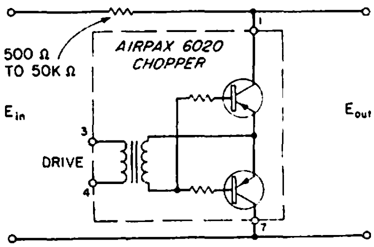


SERIES MODULATOR

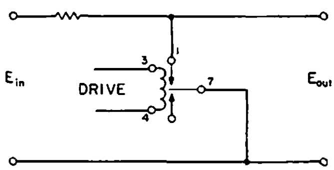


AIRPAX ELECTRO-MECHANICAL EQUIVALENT

SERIES MODULATOR: Half-wave chopping action is obtained from this circuit employing a Type 6020 transistor chopper. Signal voltage may be as high as the voltage rating of the transistor choppers. In many applications the performance is as good as with a SPDT chopper. Types 6010 and 6011 used in this application require an external drive transformer and limiting resistor. As shown, no provision is made for discharge of shunt capacitance, and high frequency performance is not good unless the output is shunted with a resistor small enough to discharge the capacitance during the "off" half-cycle.

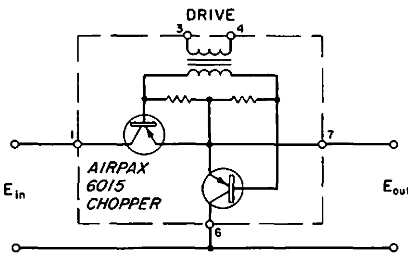


SHUNT MODULATOR

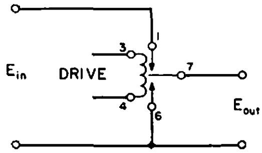


AIRPAX ELECTRO-
MECHANICAL EQUIVALENT

SHUNT MODULATOR: Type 6020 in this circuit provides half-wave chopping action. The range of the series resistance shown is representative of the chopper's "on" and "off" resistances and is of such a value as to impose no limitation on the charging of shunt capacitance. In this shunt arrangement, the discharge path is provided through the chopper during the "on" time. Types 6010 and 6011 may be used, but would require an external drive transformer and series resistor. Type 6045, designed specifically for this application, includes a 5500 ohm series resistor within the chopper enclosure.

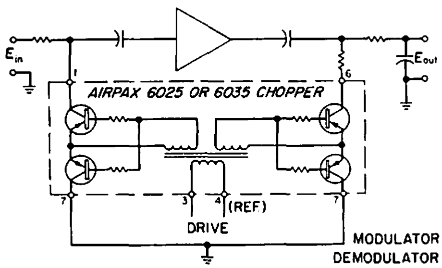


SERIES SHUNT MODULATOR

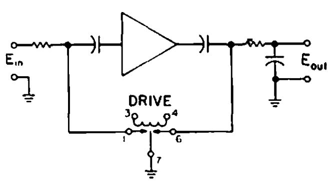


AIRPAX ELECTRO-
MECHANICAL EQUIVALENT

SERIES-SHUNT MODULATOR: In this circuit, employing a Type 6015 transistor chopper, quick charge and discharge paths are provided. Offset occurring with the 6015 in other circuits cancels out and low level operation is excellent. Even lower noise levels and minimum offset can be obtained with a Type 6010 in this circuit by replacing the resistor divider network with a potentiometer. This trims out the small differences between transistors. Types 6010 and 6015 are limited to input signals less than the driving voltage. Types 6025 and 6035 transistor choppers may be plugged directly into this type of circuit.



MODULATOR
DEMODULATOR



AIRPAX ELECTRO-
MECHANICAL EQUIVALENT

MODULATOR-DEMODULATOR: A Type 6025 or 6035 in this circuit produces half-wave chopping. Type 6015 may also be used, but greater offset and a lower maximum signal voltage limitation are introduced. Capacitive coupling, due to transistor junction capacitances between halves of the chopper, is greater than in the equivalent electromechanical chopper and may introduce difficulties in high gain circuits.

VACUUM TUBE CHOPPER DRIVERS

The compact chopper driver is a straightforward RC phase shift oscillator using one section of a dual triode combined with a power amplifier using the other section. The output at 6.8 RMS volts is developed across the secondary of an isolation transformer whose output is balanced to ground to reduce electrostatic coupling between drive coil and contacts.

These drivers are available with a dual triode and an aluminum tube shield or with a ruggedized vacuum tube and a black heat dissipating tube shield. The potted circuit within the chopper driver withstands at least as much vibration as the ruggedized vacuum tube. The seated height of the complete unit when plugged into an octal tube socket is about 3⅝″.

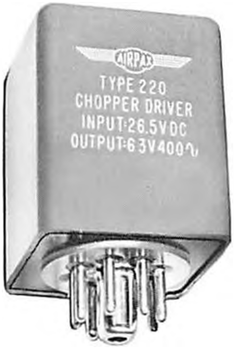
VACUUM TUBE CHOPPER DRIVERS
OUTPUT FREQUENCY 400 CPS

TYPE*	DATA NO.	PLATE VOLTAGE	OUTPUT VOLTAGE	TEMPERATURE	VACUUM TUBE
202-1	112	225	6.8±1.4VRMS	—65°+85°C	12AT7
202-2	112	225	6.8±1.4VRMS	—65°+85°C	5965
202-R1	112	225	6.8±1.4VRMS	—65°+85°C	12AT7WA
202-R2	112	225	6.8±1.4VRMS	—65°+85°C	6829
203-1	112	250	6.8±1.4VRMS	—65°+85°C	12AT7
203-2	112	250	6.8±1.4VRMS	—65°+85°C	5965
203-R1	112	250	6.8±1.4VRMS	—65°+85°C	12AT7WA
203-R2	112	250	6.8±1.4VRMS	—65°+85°C	12AT7WA
204-1	112	165	6.8±1.4VRMS	—65°+85°C	5965
204-R1	112	165	6.8±1.4VRMS	—65°+85°C	6829
205-1	112	165	6.8±1.4VRMS	—65°+85°C	5965
205-R1	112	165	6.8±1.4VRMS	—65°+85°C	6829
206-1	112	175	6.8±1.4VRMS	—65°+85°C	5965
206-2	112	175	6.8±1.4VRMS	—65°+85°C	5965
206-R1	112	175	6.8±1.4VRMS	—65°+85°C	6829
206-R2	112	175	6.8±1.4VRMS	—65°+85°C	6829
207-1	112	200	6.8±1.4VRMS	—65°+85°C	12AT7
207-2	112	200	6.8±1.4VRMS	—65°+85°C	5965
207-R1	112	200	6.8±1.4VRMS	—65°+85°C	12AT7WA
207-R2	112	200	6.8±1.4VRMS	—65°+85°C	6829

*Suffix number indicates: —1 operates one Airpax 300 chopper; —2 operates two Airpax 300 choppers; —R is ruggedized. Types 204-1 and 204-R1 are 480 CPS chopper drivers.

Filament Voltage: Either 6.3 volts at 0.3 ampere or 12.6 volts at 0.15 ampere. AC or DC will operate the vacuum-tube heater.

TRANSISTORIZED CHOPPER DRIVERS



Chopper Drivers Types 220 and 221, are capable of supplying power for one or two Airpax miniature choppers. The chopper driver output is approximately 6.5 volts at 400 CPS, the input is 24 volts DC. Two outputs are provided for operation of either one or two choppers. The 220 is a plug-in component, the 221 is its equivalent in a bolt-down package. There is no electrical difference.

Types 220 and 221 chopper drivers are potted and sealed in a steel can 1½″ square with a 2 1/16″ maximum seated height. The Type 220 plugs into a standard octal socket. Weight is approximately 8 oz.

TRANSISTORIZED CHOPPER DRIVERS

TYPE	DRIVE	VOLTAGE OUTPUT	OUTPUT FREQ. CPS	TEMPERATURE	VIBRATION
220	24V DC	6.5±0.5VRMS	400	−40°+72°C	10G 55 to 2000 CPS
221	24V DC	6.5±0.5VRMS	400	−40°+72°C	10G 55 to 2000 CPS

NOTES: (1) Transistorized chopper drivers will drive either 1 or 2 choppers. (2) Type 221 is a bolt-down unit with solder terminals.

ELECTRICAL CHARACTERISTICS

INPUT: 20 to 30 VDC at approximately 50 MA, 26.5 V nominal.

OUTPUT VOLTAGE: 6.5 V±.5 volts at 24 VDC input.

OUTPUT FREQUENCY: Nominal 400 CPS to be within ±20 CPS under any normally encountered set of environmental conditions with a two chopper load.

WAVE FORM: The output is essentially a sine wave.

STARTING VOLTAGE: 18 VDC max.

HIPOT TEST: Between any input terminal and case or ground. 500 RMS volts at 60 CPS for 1 minute.

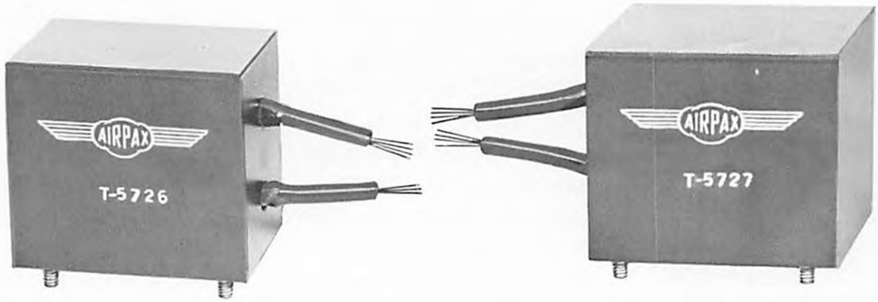
INSULATION RESISTANCE: At least 10 megohms at 500 VDC between any input terminal and case or ground.

ELECTROLYSIS: Minimum of 1000 hours at 72°C and 100 VDC applied between any input terminal and case or ground.

ISOLATION: Input terminals are isolated from case and ground. Output is connected to case and ground at transformer center tap.

Reliable operation is achieved when mounted in any position and operated under natural combinations of normally encountered environmental conditions.

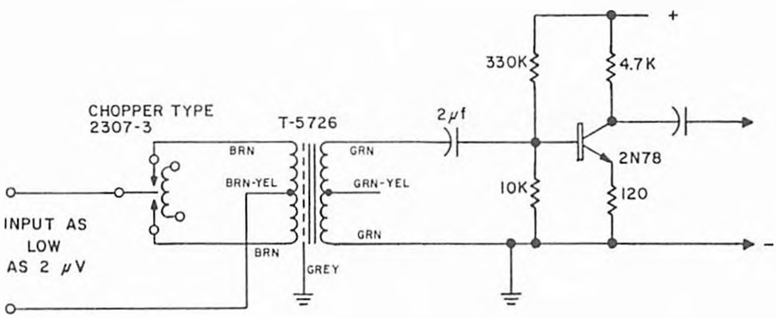
TRANSFORMERS FOR CHOPPER CIRCUITS

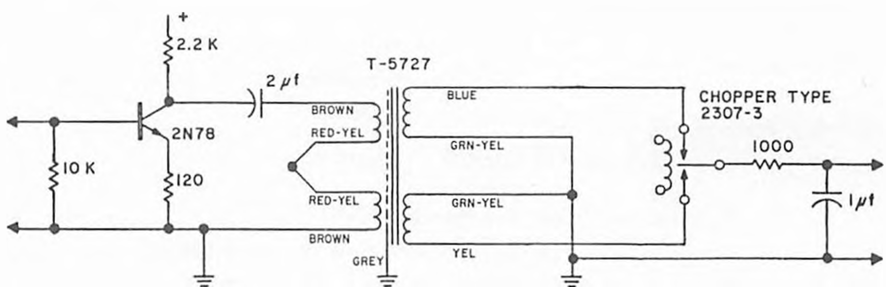


INPUT AND OUTPUT CHOPPER TRANSFORMERS

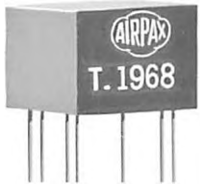
TYPE T-5726 input chopper transformer has a center-tapped primary, 400 ohms each half, and a 4000 ohm secondary (total), also center-tapped. It is designed for use with a 400 CPS low noise chopper. Construction includes nested hum shields (mu-metal and copper), and balanced windings. The connections are brought out as copper leads to prevent thermal junctions.

TYPE T-5727 output transformer is similar in constructional features to Type T-5726 but has a 3000 ohm center-tapped primary working into a 300,000 ohm center-tapped secondary.





DRIVE TRANSFORMERS FOR TRANSISTOR CHOPPERS



TYPE T-1817-1 is a miniature chopper drive transformer designed to drive a single Type 6010 or Type 6011 transistor chopper.

TYPE T-1904-1 is designed to drive two Type 6010 or Type 6011 transistor choppers. Same size as T-1817-1 above. Has 2 secondaries.

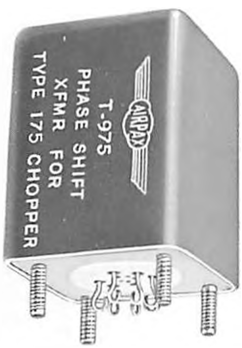
TYPE T-1968 is a fully encapsulated miniature transformer for printed circuit applications. It is designed to drive two Type 6010 or Type 6011 choppers.

TRANSISTOR CHOPPER DRIVE TRANSFORMERS

MAX. AC DRIVE VOLTS: TYPE T-1817-1, 7 V. at 50 CPS, 14 V. from 100 CPS up; TYPES T-1904-1 and T-1968, 10 V. from 400 CPS up.

TYPE	FREQ. RESPONSE	TURNS RATIO	DIMENSIONS L	W	H
T-1817-1	50 CPS to 5 KC	1:1	1 7/32	21/32	19/32
T-1904-1	200 CPS to 20 KC	1:1:1	1 7/32	21/32	19/32
T-1968	200 CPS to 20 KC	1:1:1	3/4	5/8	1/2

Types T-1817-1 and T-1904-1 have channel frame strap mounting. Type T-1968 has printed circuit leads (100 mil. grid spacing). All are high pot tested at 500 volts.



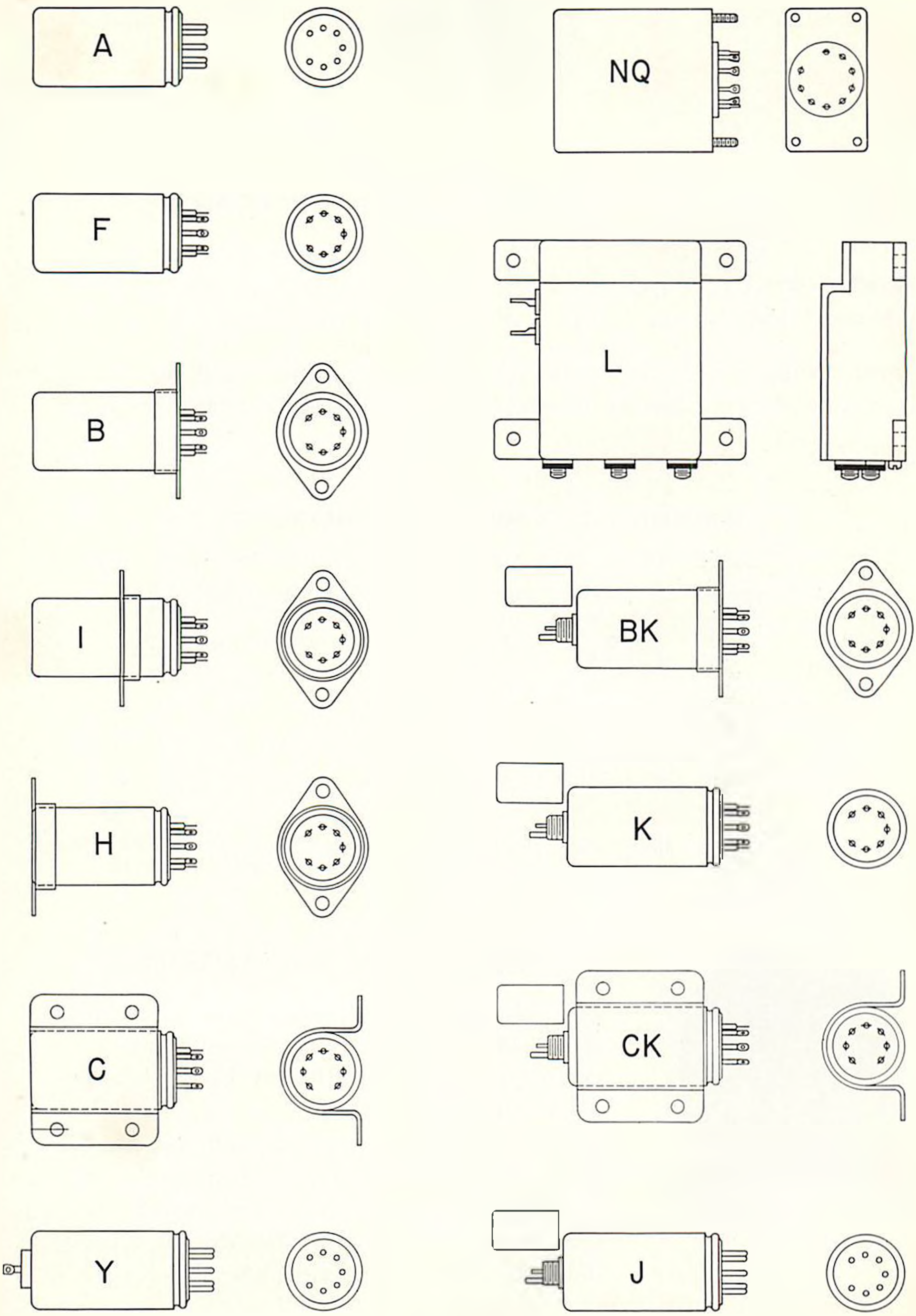
PHASE SHIFT TRANSFORMER

The phase lag between contact action and driving signal, which may be between 15 and 80 electrical degrees, can be reduced to zero by the use of a phase-shift transformer.

TYPE T-975 is an Airpax phase-shift transformer designed for use with a Type 175 chopper operating from a 60-cycle source. The center-tapped secondary permits a drive balanced to ground, thus keeping noise at a minimum.

CHOPPER MOUNTING STYLES

For mounting dimensions see specification covering individual type.



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