CIRCUITS
- Electronic Oscillators
- Regulated Power Supply Design
- Ferroresonant Circuits

TV CIRCUITS AND APPLICATIONS
- Television Inter-Imagery Sound Reception
- Video IF Design
- Cathode-Ray High-Voltage Supplies
- Television Interference Filters
- Television Reception at "Shadows" Locations

WAVE FORMS AND WAVE SHAPING
- Non-Sinusoidal Wave Forms
- Audio Frequency Distortion Measurements

METERS AND MEASUREMENTS
- The Direct-Current Meter
- High Resistance Non-Electronic, D.C. Voltmeters
- Application of the Electrometer
- Improved, Crystal-Type Noise Generator

COMPONENTS
- Fixed Capacitors in Modern Circuits
- Proper Use of By-pass Condensers
- Non-Linear Resistors
- Amateur Applications of Crystal Diodes
- Proper Electronic Wiring Techniques

SPECIALIZED APPLICATION AND DEVELOPMENT
- Using Standard Time and Frequency Broadcasts
- Photoelectric Cell Applications
- Printed Electronic Circuits
- The Transistor, An Amplifying Crystal
- Junction Transistor Circuits
- Class-B Transistor Amplifier Data
- Loud Lunes in Transistor Amplifier Design
- Simple, Inexpensive Geiger Counters
- Recent Trends in Single-Sideband Communication
- The Citizens Radio Service
- Elementary Binary Arithmetic
# TABLE OF CONTENTS

<table>
<thead>
<tr>
<th>Section</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>I TV Circuit and Applications</td>
<td>1</td>
</tr>
<tr>
<td>II Circuits</td>
<td>15</td>
</tr>
<tr>
<td>III Components</td>
<td>39</td>
</tr>
<tr>
<td>IV Wave Forms and Wave Shaping</td>
<td>55</td>
</tr>
<tr>
<td>V Meters and Measurements</td>
<td>67</td>
</tr>
<tr>
<td>VI Specialized Application and Development</td>
<td>85</td>
</tr>
</tbody>
</table>

INDEX | 119 |
SECTION I
TV CIRCUITS AND APPLICATIONS

Television Inter-Carrier Sound Reception

In all present-day television systems, the sound, picture and synchronization information are transmitted simultaneously over a frequency-scrambled television channel. Prior to 1943, this was accomplished solely by the use of an amplitude-modulation (AM) method, in which the audio intelligence and the video intelligence were each impressed on two completely separate carriers which differed in frequency by 4.5 megacycles. Subsequent developments in the art demonstrated the desirability of utilizing the much-advanced fidelity and noise reduction properties inherent in frequency-modulation (FM) to improve the audio performance of the television system, and as a direct consequence to increase listener enjoyment and acceptance of the new mode of entertainment. This combination of AM for the video carrier and FM for the sound carrier, with transmission standards as established by the F. C. C., is currently in use by all television broadcasting stations in this country. This changeover from AM audio to FM sound while requiring rather extensive modifications of the transmitter, did not involve obliteration of existing TV receivers, since the FM sound could be received by the slope detection method. Fig. 1 graphically illustrates a typical reception analysis for one of the 12 U.S. channels in use at the present time.

In recent years, a number of schemes have been disclosed for the simultaneous transmission of both picture and sound on a single carrier. In general these systems were based on the use of a multiplex or time-division method of transmission, in which it was proposed that the audio modulation be impressed on the video carrier during the intervals normally reserved for the synchronization pulses. Because of the fact that these sync pulses are absent during the intervals when the synchronizing pulses are present, it is evident that adoption of such a system would not only require a major modification of the transmitters but also make existing receivers obsolete.

In 1947, exposition was made in papers by L. W. Parker, and R. B. Dome, of the development of a new method for recovery of the audio intelligence contained in a composite television signal. In this system, a 4.5 Mc. best frequency, usually present as an undesired signal in the control grid circuit of the kinescope, is modulated to produce a signal which can be amplified, clipped or limited and then demodulated in the usual manner.

That such a 4.5 Mc. signal can exist may readily be seen from the following generalized consideration. It may be demonstrated in passing that even though this ordinarily undesired signal is usually present in greater or lesser degree in all TV receivers, only one manufacturer, to the writer's knowledge, has made particular provision for 4.5 Mc. trapping in the video-amplifier circuit.

It will be remembered that in a conventional superhet-type receiver, the incoming signal of frequency F and a locally generated oscillator signal of frequency f are both passed through some nonlinear device, variously termed the mixer or converter or first detector. In the conventional receiver the two signals are combined and form an IF signal of frequency equal to the difference between F and f. It must be stressed at this point however that this heterodyne or best frequency will be generated only if the mixing
device (vacuum tube, crystal diode, thermistor, etc.) possesses some non-linearity in its impedance characteristic.

In the conventional dual i.f. television receiver, both the picture and sound carriers, comprising the composite television signal, are amplified by a common broad-band r.f. stage, heterodyned with a single local oscillator signal and converted to a complex signal containing video and audio i.f. components. Although the conversion process loses the frequency of the two carriers, the 4.5 Mc. frequency difference between them remains unaltered. If, as is usually the case, the oscillator is operated on the high-frequency side of the signal, a side-band reversal takes place, i.e., the relative positions of the video and audio carriers are interchanged.

The use of a common i.f. amplifier, mixer and oscillator for both picture and sound adds to ease of operation and at the same time affords a material saving in cost. A natural extension of this dual function technique to the i.f. portion of the receiver has resulted in the recent appearance on the market of several commercially designed TV receivers in which one or even both of the i.f. stages are made to function as combined picture and sound mixer. From one point of view, however, two signals are separated by approximate filters or traps, amplified further as required and then demodulated, as shown in Fig. 2.

In the interferor sound system, Fig. 3, both the picture and sound i.f. signals are handled simultaneously by a common wide band i.f. amplifier. In the video detector, usually a vacuum tube or germanium crystal diode, the frequency modulated sound i.f. signal is heterodyned with the amplitude modulated video i.f. signal. The resulting 4.5 Mc. beat frequency, produced by the non-linearity of the detector characteristic, is both frequency modulated by the sound and amplitude modulated by the video. It has been shown, however, that if a low level FM signal is heterodyned with a high level AM signal, the resultant signal is largely frequency modulated and is relatively free of AM. This 4.5 Mc. signal is further amplified by the video amplifier, passed through one or more amplifier stages, and is finally demodulated by any of the well-known methods of FM detection. Use of the ratio detector for this function provides adequate AM clipping without resort to a separate limiter.

Although not adequately demonstrated by the simple block-diagrams of Figs. 2 and 3, the interference-type system presents a considerable simplification in circuitry when compared with the actual schematic of the dual i.f. system of television reception. In addition to this advantage, with its attendant economy, the Pakeker system possesses several other desirable characteristics among which may be mentioned simplicity of tuning, freedom from oscillator drift and microphonics, and reduction of inter-channel crosstalk.

The dual i.f. television receiver suffers from the serious disadvantage that even comparatively slight mismatching or drift of the local oscillator seriously affects both audio quality and discriminator amplitude-saturation susceptibility. In common with other types of AFC receiver, background noise and hiss are also increased by such oscillator malfunction. Automatic frequency control of the oscillator has been applied in certain television models as a means of reducing these shortcomings, but this of course adds to the complexity and consequently to the cost. A commentary on the efficacy of AFC circuits as applied to TV receivers may be made by noting that one manufacturer, after marketing a receiver using AFC, subsequently issued a modification kit for adding a fine tuning control to sets in the field.

These conditions are further aggravated by the fact that the TV sound deviation is only 12.5 kc as compared with 37.5 kc for standard FM broadcast transmission. The discriminator output can therefore usually be made higher, requiring almost pinpoint accuracy in TV set tuning.

Sixty-cycle frequency modulation of the local oscillator in the standard dual i.f. system, due to insufficient filtering or inadequate linear leakage, is often manifested as objectionable hum in the audio. To eliminate this undesirable effect, a side-band rejection filter or input stage must be provided to reduce such hum. Acoustical feedback from the speaker of a dielectric system is any microphonic portion of the oscillator circuit, such as a trimmer condenser plate, an inter-winding capacity of the elements, and the like, can cause annoyance or even audio level as high as 60 db.

The Parker System, in which ease of tuning is an advantage of the 4.5 Mc. frequency difference between carriers, which is used as the sound channel, is maintained by accurate control at the transmitter rather than by the relationship existing between r.t.
ceived and locally generated signals, is immune to these troubles. Since the video i.f. is normally of the wide band type, misadjustment at drift of the local oscillator sufficient to cause severe distortion of the sound in dual-i.f. receivers, causes no appreciable degradation of either picture or sound quality in sets equipped for intercarrier sound reception.

The discussion thus far of the comparison between the two systems has been confined to the credit side of the intercarrier ledger. There are, of course, the ever present debits. Chief among these is the susceptibility of the inter-carrier system to audio interference caused by: (a) frequency or phase modulation of the video carrier, (b) momentarily disengagement of the video carrier during modulation peaks (c) failure at the transmitter to accurately maintain the prescribed 4.5 Mc difference between video and audio carriers, and (d) drift of the receiver discriminator tuned circuits.

The remedial measures necessary to reduce the above effects present no insurmountable problems, as may be seen from the following considerations:

(a) Frequency or phase modulation of the video carrier produces, in the Parker system, undesirable modulation and distortion products in the reproduced sound. This can be prevented, or at least minimized, by proper transmitter design and adjustment.

(b) Since the inter-carrier system depends upon both carriers in sound reproduction, it is evident that momentary disengagement of the video carrier, such as might be caused by over-modulation of the transmitter, also causes interruption of the sound. By imposing the limitation at the transmitter that the picture carrier shall never fall below 10 or 15% of maximum amplitude, the possibility of such sound "break-up" is prevented. Another effect which has symptoms very similar to sound break-up in the video carrier disengagement is that occasioned when one of the video i.f. amplifier stages is driven to cutoff, as it is sometimes possible to do by improper adjustment of the contrast control. Since the peaks signals are the "even" pulses, which are in the infra-black region, this type of operation is not necessarily deleterious to the performance of the dual-i.f. sets. In the Parker system, however, operation at cutoff results in sound break-up.

(c) Failure at the transmitter to accurately maintain the 4.5 Mc spe- cification of the video and audio carriers results in the same kind of audio performance degradation as that caused by misadjustment of the tuning control or discriminator circuit in the conventional dual-i.f. system. Where- as this condition can easily be cor- rected by the user of the older system, by adjustment of the fine-tuning control, the inter-carrier arrangement must depend upon the broadcaster for the correction of this condition.

(d) The effects of drift in the dis- crimator circuit of the inter-carrier sound system are usually slight since the sound sub-i.f. is at a relatively low frequency and can be virtually eliminated by careful design and the judicious use of small fixed capacitors having the proper capacitance versus temperature characteristic.

Figs. 4 and 5 show the overall re- sponse curves for the two TV systems discussed above. The only difference is in the shape of the lower curves in the vicinity of the sound portion of the spectrum. In the dual-i.f. system, full response over at least a 25 kc. bandwidth, centered on the sound "testing" frequency is desired, while in the inter-carrier method the sound acceptance notch is in the form of a small shelf of similar width and cen- ter frequency. The reason for a plat- eau rather than a gradual roll-off for the intercarrier sound response is that in this way partial demodulation of the FM sound by slope detection and possible picture interference is prevented. As shown by Fig. 3, this response at the sound shelf shall not exceed 5% of full amplitude. This, as mentioned earlier, reduces the video amplitude modulation of the sound sub-i.f. to a negligible value and also prevents possible distortion of the picture by the sound modulation.

The sound trapping requirements in the Parker system are not nearly so stringent as in conventional sets. A single trap, providing about 26 db. of attenuation is required immediately preceding the video detector to de- crease the sound response to the 5% level. This rather large attenuation is subsequently compensated for by the additional gain of the one or two video amplifier stages.

In closing, it may be recalled that the results of rather extensive field tests, and the likelihood of industry- F. C. C. cooperation in setting up the transmission standards necessary for uniformity of performance, should make the system well lead to the eventual com- plete replacement of the Parker system for television sound reception.
Video I. F. Amplifier Design

In modern radio communication and television sets, the necessity of transmitting and receiving a large amount of intelligence per unit time, or of handling wave forms which contain high frequency components, imposes difficult requirements upon the bandwidth of the circuit used. In the receiver system, for instance, the modulation of the transmitted signal by very short, rectangular pulses of energy, results in the i.f. output occupying a broad band or spectrum of frequencies. The width in megacycles of the band required for the transmission of such rectangular pulse signals is expressed to a rough approximation by:

\[
\text{Bandwidth (MHz)} = \frac{2}{\text{rise time (microseconds)}}
\]

Thus, a radar transmitter being modulated by 5 microsecond pulses would occupy a band (exclusive of minor side bands) of 2 divided by 5 megacycles. In television, the transmission of high-definition picture information consisting of several million elements per second, as well as synchronizing pulses and sound, requires the allocation of a 6 megacycle channel for each transmitter in operation. In any such broad bandwidth system, if the receiver is to recover as much of the transmitted signal as possible, it must be capable of simultaneously amplifying the entire band of frequencies transmitted and amplifying each equally. In the superheterodyne type of receiver, the saturation of this requirement greatly affects the design of the i.f. amplifier, since it is this channel of the receiver which determines the overall selectivity to a large extent.

Fortunately, the design of a high-band or "video" intermediate-frequency amplifier has been greatly simplified by recent research work. As a result, the design of high gain amplifiers capable of essentially "flat" bandwidth characteristics as wide as 10 megacycles is relatively uncomplicated.

The bandwidth of an i.f. amplifier is taken as the frequency difference between points 3 db. down from the maximum amplitude on each side of the response curve and is symbolized by \( \Delta f \). See Fig. 1. In the simplest form of amplifier stage, which is the single-tuned circuit shown in Fig. 2, the bandwidth in megacycles is given by:

\[
\text{Bandwidth (MHz)} = \frac{1}{2 \pi \text{RC}}
\]

As this relation shows, the bandwidth of a single-tuned stage is inversely proportional to both the shunt capacity and the shunt resistance. In practice it is the resistance which is varied to control the slope of the response curve. The addition of "loading resistors" across the tuned circuits, common in television and other video i.f. circuits, broadens the response as illustrated by the dotted curve in Fig. 1. Loading the resonant circuit in this manner, of course, reduces the maximum response or gain at which is shown. The bandwidth at the new 3 db. point has been increased, but the peak response has been sacrificed proportionately in favor of bandwidth. This demonstrates the important fact that the gain-bandwidth product of such an amplifier is a constant. This means that a stage giving a gain of 10 over a bandwidth of 10 megacycles may also be made to deliver a gain of 5 at a 2 megacycle band-pass, or any other combination whose gain-bandwidth product (G x B) is equal to ten. The gain-bandwidth product, which is the accepted "figure of merit" of an amplifier stage, depends on the transconductance (g_m) of the tube type used and the total distributed shunt capacity in the following manner:

\[
g_m \times B = \frac{1}{2 \pi C}
\]

Since the gain-bandwidth product is inversely proportional to \( C \), which includes the distributed wiring capacity as well as the tube interelectrode capacitance appearing across it, it is very important in circuit layout to reduce stray capacity to a minimum. In practical circuits using modern tubes, the total \( C \) may be limited to 10 mmf. Table 1 shows the Gm products for some frequently used tubes, allowing 5 mmf. for distributed circuit capacity.

<table>
<thead>
<tr>
<th>Tube Type</th>
<th>Gain-bandwidth Product (MmF)</th>
</tr>
</thead>
<tbody>
<tr>
<td>6AX7</td>
<td>4000</td>
</tr>
<tr>
<td>6A86</td>
<td>4000</td>
</tr>
<tr>
<td>6A95</td>
<td>5000</td>
</tr>
<tr>
<td>6A45</td>
<td>5000</td>
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</table>

Table 1

Unfortunately, when single-tuned amplifier stages are connected to the same frequency (synchronously tuned) are decoupled, the overall band-pass does not remain that of the individual stages, but is reduced radically with the number of stages. Four stages,
each 4 megacycles broad at the 3 db point, when cascaded would thus have an overall band-pass of only 1.75 megacycles. This is evident from the fact that if the voltage gain at the center frequency (fo) is 10, the gain at the 3 db points is only 7.07. Upon amplification by a second identical stage, the gain at fo is 10 x 10 or 100 while the gain at the former 3 db points is now only 7.07 x 7.07 or 50, which is 6 db down in voltage. The bandwidth at the 3 db point has been reduced to 64% of that for the single stage. Further amplification by similar stages would result in the overall bandwidth being reduced to 51% for a third stage, 44% for a fourth stage, 20% for the fifth, etc.

In addition to the undesirable feature of rapidly decreasing pass-band for multiple stages, the synchronously-single tuned system does not satisfy the requirements of the television video i.f. since it is incapable of producing the flat-topped response curve required for picture reproduction. The shape of the video i.f. response which is accepted as the standard in television practice is shown in Fig. 3. An essentially flat, band-pass of nearly 4 megacycles is required for high-definition picture reproduction on large-screen cathode-ray tubes, although sets using small tubes may get along with much less. The gradual, nearly linear decrease in the response at the picture-carry end of the curve is important in avoiding the presence in the transmitted signal of the first 1.25 mc of the lower side-band. (The 20 db is supposed at the transmitter). When the picture-carry i.f. frequency is aligned to the mid-point of this slope, the small portion of the vestigial lower side-band which is under the response curve is cancelled by the response to a similar area from the lower 1.25 mc of the upper side-band. Therefore, the response to the lower side frequencies is made nearly equal to the horizontal though derived part of the both upper and (vestigial) lower transmitted side-bands.

Considerable improvement over the performance of synchronous single-tuned amplifiers may be obtained by the use of multi-tuned circuits. In a multi-tuned, transformer-coupled stage such as is shown in Fig. 4, the coefficient of coupling (x) between the primary and secondary circuit Q's may be adjusted so that the response curve is essentially flat-topped. Thus, if a maximally flat or "transitional" coupling occurs when the circuit Q's are the coefficient of coupling are related as shown in Fig. 4. The term "transitional coupling" is derived from the fact that the coupling is adjusted to the point of transition between the single and double-humped response curve. It will be recalled that, as the coupling coefficient of the tuned transformer is increased from a very small value, the curve of secondary current versus frequency changes from a small sharp peak when the circuits are under-coupled, to a broad double-peaked response when the circuits are over-coupled. (Dotted lines, Fig. 4). The coefficient of coupling of the inter-stage transformer may then be determined by measuring the capacity values necessary to resonate the primary to a given frequency when the secondary is alternately open- and short-circuited. (Cs and Cs respectively.) Knowing the ratio of these capacities:

\[
\text{Coefficient of coupling } (x) = \frac{1}{\sqrt{2}} \times \frac{C_2}{C_1}
\]

At the value of x corresponding to the critical coupling, the transfer of energy from the primary to the secondary is maximum and the curve is the "flat" type. The response characteristic obtained in this manner is more nearly that required by the television video i.f. Furthermore, because of the more uniform response over the pass-band, the overall band-width does not decrease as rapidly when identical Q's are cascaded in the case of synchronously single-tuned stages. When two double-humped, translationally-coupled amplified stages are cascaded, the output bandwidth is reduced to 80% of the width of the individual stage. The corresponding figure for synchronous single-tuned stages is 64%.

![Fig. 4](image)

Further improvement in pass-band width performance may be attained by the use of more complicated inter-stage coupling networks. These include, double-tuned staggered, triple-tuned transformer-coupled, single-tuned inverse-feedback and complex filter-coupled stages. Most of these types are difficult to design and troublesome to construct and align, so will not be discussed here in detail.

One type of band-pass amplifier which does retain the simplicity of design and alignment of the synchronously single-tuned type, and yet overcomes most of its disadvantages, exists in the staggered amplifier. Walman and others have shown that if the successive stages of a simple single-tuned amplifier are adjusted to slightly different frequencies (staggered) throughout the desired pass-band, the composite response curve may be made flat-topped and the gain high. Furthermore, the design work requires only high school math and a few simple tables, the construction done with common tools and the alignment may be accomplished in a few minutes with the aid of a spot-frequency signal generator and an output meter. The double-tuned and other more complex types previously mentioned require the use of a sweep-frequency signal generator and an oscilloscope. Staggered-tuned systems are being used extensively in commercial television practice.

Since the individual stages of the staggered amplifier are nearly the single-tuned type shown in Fig. 2, the design equations (2) and (3)
<table>
<thead>
<tr>
<th>TABLE II</th>
</tr>
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The widespread use of the cathode-ray tube in television viewers and in test instruments has occasioned the use of high-voltage direct-current supplies of rather unconventional design. Such departures from the standard 60-cycle power supply are due to the special requirements of the tube itself, for operating voltages ranging up to 30,000 volts at current drawn less than one milliampere. The production of such voltages by line-frequency iron-core transformers is impractical because of the process of winding high-voltage secondaries with extremely fine wire is expensive. The problem of maintaining adequate insulation between the secondary and the other parts of the transformer is also troublesome. In addition, the network required to filter the low ripple-frequency is dangerous and bulky. For these reasons, the use of this type of supply for high potentials has been virtually superseded by the more modern high-voltage supplies.

There are three general types of special high-voltage, low-current supplies in common use for cathode-ray tube applications. In the order of popularity, they are:

- **Stagger-tuning table**: A Q Meter where available, or by empirical formula. Since additional capacitance is very detrimental to the bandwidth, it is essential that the circuit be matched to the circuit capacity or tuned with high-quality power-supply filters.

When resistors and inductors corresponding to the values determined for R and L are incorporated in a single-tuned stage such as that shown in Fig. 2, and the stages are connected in cascade, the resulting stagger-tuned amplifier is non-critical to adjust and will compare favorably with more complex types in performance. The overall gain-bandwidth product is better than a synchronously tuned amplifier of the same number of stages by a factor greater than two. Alignment is accomplished by connecting a standard AM signal generator to the input of the amplifier and an amplitude indicating device such as a voltmeter to the output. The signal generator may then be set to the recommended stagger frequencies in succession and the individual stages corresponding to this frequency peak for maximum output response. Due to the isolation action of the tube, there is virtually no interaction between stages while tuning.

Cathode-Ray Tube High-Voltage Supplies

---

*Note: The table and text are from a technical or engineering perspective, detailing the design and use of cathode-ray tubes in television applications.*
A typical high-voltage supply of the flyback, or "kick," variety is shown schematically in Fig. 1. This circuit makes use of the fact that a high-voltage pulse of considerable duration is developed across the primary winding of the horizontal deflection transformer during the flyback period of the saw-tooth current pulse which is generated on the plate circuit of the horizontal amplifier tube. The amplitude of this inductive voltage surge is expressed approximately by:

\[ V_{\text{out}} = L \frac{dI}{dt} \]

where: \( V_{\text{out}} \) is the primary winding induction.

Thus, it is seen that the amplitude of the voltage "kick" developed depends upon the inductance of the transformer primary winding and the rapidity with which the current flowing through it is changing. Since the current in the transformer builds up during the sweep period of about 57 microseconds, and collapses rapidly during the flyback time of approximately 7 microseconds, the rate of current change, and hence the induced voltage, is more than eight times greater during the retrace period.

This inductive voltage "kick" is further stepped-up by a third winding on the horizontal deflection transformer which is connected with the primary to form an auto transformer. In addition to the high-voltage rectifier and filter circuit shown in Fig. 1, this special horizontal deflection transformer having a tertiary winding for the high-voltage supply is the only extra component added for the flyback supply, since all other parts are already standard components of the horizontal deflection system. The transformer, T1 in Fig. 1, is of special pole-transformer construction. The windings are assembled on a core built up of very thin laminations or a special low-loss molded iron powder core.

The Rectifier and Filter Circuit

The type of high-voltage rectifier and direct-current filter used with the flyback supply is also practically standard for the other types of high-voltage supplies to be discussed. For this reason, these components will be discussed in the detail below.

To rectify the high-voltage alternating or pulse waves, a specially designed pulse rectifier tube is almost universally used. This tube, designated as type 6AK5, is rated at 33 kilovolts maximum peak inverse voltage. Maximum diode current is rated at 2.5 milliamperes. The specified characteristic requires only 1.2 watts at 250 milliamperes for heating. This low filament power consumption feature of the 8016 rectifier, (less than one-quarter watt) enables the tube to be heated directly from the source of high-frequency energy. The filament circuit is inductively coupled to the high-voltage step-up transformer by means of a one- or two-turn link, as indicated in Fig. 1. A 2.3 ohm resistor is sometimes used in series with the filament circuit to protect the tube from burnout in case of over-coupling. The low-voltage link coupling feature of the 8016 facilitates isolating the rectifier filament circuit, which is operated at full output voltage above ground.

The design of the smoothing filter used with all modern types of high-frequency supplies is greatly simplified because of the low current requirement and the high ripple-frequency. In the case of the tube-type supplies, the sawtooth frequency is the same as the horizontal sweep frequency so that the tube filament, in the r.f. supply, the frequency of operating is usually above 50 k. When such high-frequency waves are enclosed, the resulting ripple can be filtered by relatively low-cost components. The high capacity filter capacitors and heavy iron-core chokes necessary in the uses of a low-frequency filter section are replaced by a simple high-frequency filter consisting of a 0.0005-microfarad condenser and a low-voltage resistor. The output capacitance of this filter, shown by dotted lines in Fig. 1,
1. frequency consists only of the condenser-to-lead capacity of the high-voltage output cable.

Because of the low storage capacity of this filter design, the high-frequency supply is much less dangerous to operating personnel, than a line-frequency supply using a "brute force" filter.

Voltage doubling and tripling arrangements are frequently resorted to in high-frequency supplies to obtain higher potentials, or to provide convenient taps for intermediate voltages. Fig. 2 illustrates some representative voltage-multiplying circuits. It will be noted that the techniques used are similar to those practiced in conventional line-frequency power supplies.

The Radio-Frequency Supply

The basic circuit of the radio-frequency type of high-voltage supply is shown in Fig. 3. In this arrangement, the r.f. output voltage of a class C self-oscillating oscillator is transformed to a very high voltage by a tightly coupled, double-tuned, step-up transformer. Since the frequency of operation may be between 50 to 200 kilocycles per second, the voltage step-up is accomplished by a lightweight, coil-wound, air-wound transformer.

The power oscillator usually consists of a beam pentode of the 6V6 or 6L6 type used in a tuned-plate, grid-feedback circuit. Class C operation is used for high efficiency. Anode voltage ranging from 250 to 400 volts d.c. is applied to the oscillator tube, depending on the type used and the high-voltage required. In applications where rectifying potentials as high as 27,000 volts are required for projection television kinescopes, two or more oscillator tubes as large as the type 807 may be operated in parallel.

The step-up transformer which couples the r.f. oscillator to the rectifier and filter circuit is designed to fulfill several special requirements. It must be sufficiently insulated between the primary and secondary windings to withstand the full output voltage with the distributed capacity of the wiring and rectifier tube. Wiring for the control grid and grid-feedback "tinkler" must also be provided.

Fig. 4 shows a typical r.f. supply transformer design. The coil form is usually of thin-walled, impregnated bakelite tubing of low power factor. To decrease leakage currents between windings, and to facilitate the free circulation of air for cooling, a series of long circumferential slots is made in the coil form, as shown in Fig. 4. The primary, secondary, and grid-winder windings are wound with Litz wire to minimize losses. The high-voltage secondary winding is made of universal-wound "film", spaced sufficiently to prevent corona discharge between them.

The overall efficiency of the r.f. oscillator type of high-voltage supply is between 25 and 45 percent, its principle disadvantage lies in the fact that sufficient harmonic radiation is sometimes present to cause interference with other circuits. Complete shielding and supply-leaf filtering is therefore necessary.

The Pulse-Type Supply

Although not enjoying the wide usage of the high-voltage supply types discussed above, the pulse-type supply is an accepted practice. Like the flyback supply, the pulse-generating circuit is synchronized to occur during the blanking period of the horizontal sweep cycle, so that minimum radiation interference is caused in many reception sets. In connection with the r.f. supply, it has the disadvantage of requiring additional components and power, but can be used for cathode-ray tubes having electrostatic deflection.

The circuit arrangement of the pulse-type high-voltage supply is illustrated in Fig. 5. The essential parts consist of a blocking oscillator pulse generator, a pulse amplifier stage, a step-up pulse transformer, and a rectifier-filter section similar to those used with supplies of the r.f. type.
and "kick" types. As is the "kick" supply, the amplitude of the high-voltage developed is dependent upon the rapid change of current flowing in the pulse transformer inductance during the pulse.

The pulse generated by the blocking oscillator is applied to the horizontal sweep so that it occurs during the 7 microsecond recovery period. Synchronization is effected by injecting horizontal sweep voltage into the blocking oscillator circuit at the point marked "sync" in Fig. 5. The synchronizing voltage permits oscillation of the blocking oscillator, which is held normally inoperative by a bias voltage developed across the cathode resistor, R1. In addition to preventing picture tube "blanking" by resetting the high-voltage pulse to the "dark," recovery period of the sweep, this system of synchronization also protects the kinescope from screen burning by holding the high-voltage supply inactive in case of sweep failure.

Service Notes
In performing any operation on active high-voltage supply circuits, it must be remembered that LETHAL VOLTAGES ARE PRESENT. Although the poor over-load regulation of the high-voltage types renders them much less dangerous than the older 60-cycle supplies, direct contact with the voltages developed is extremely painful and can be fatal. Therefore, due caution should be exercised.

The trouble most frequently encountered with high-potential supplies is current discharge from the high-voltage portion of the circuit. Current, identified by a blue glow of finish discharge around the parts affected, causes erratic output voltage with attendant poor picture brilliance. It is caused by high voltage gradients between adjacent parts, resulting in ionization of the surrounding air. To reduce these effects, commercial supplies are designed with all components and conductors having large radial contours since sharp corners or points of curvature attract discharge. For this reason, when working on high-voltage sources, care should be taken to avoid introducing high gradient points such as rough, sharp solder points or sharp bands in wiring. Actual current leakage on the surface or through insulation is another problem in cathode-ray tube supplies. This effect tends to load the supply excessively and frequently results in complete breakdown and carbonization of the leak path. In such cases, replacement of the defective part is the only effective remedy. Leakage can be reduced by preventing the formation of grease, dust or moisture on the surface of insulating material.

**Television Interference Filters**

The current of television broadcast has brought about many new problems in interference elimination. Much of this interference is caused by spurious radiations from transmitters of other services. The burden of finding solutions is such great that it is practically impossible for the transmitter causing the interference, and upon the owner of the set being plagued with — or his service technician. Usually, a satisfactory solution can only be arrived at through the complete cooperation of all parties concerned.

The American amateur radio operator, because of greater numbers, closer proximity to owners of TV sets, has spear-headed the technical battle to find cures for this threat to his hobby. Now, with many "hands" again able to operate at full one-kilo- watt limit in the midst of dozens of TV receivers, the battle has been won. There remain only the job of educating others in the methods employed.

The most powerful tool which has been applied to the elimination of television interference (TVI) is the frequency selective filter. The application of filter networks to television interference elimination and the construction of practical filters for use at the source of the interference, as well as at the TV set, will be discussed here.

**Causes of Interference**

Because of the lack of selectivity inherent in modern television receivers, they are particular prone to interference by spurious signals of many kinds. When one considers that the minimum band-pass for tuned receiver "front-end" is 100,000 cycles, and that many using untuned grounded-grid r-f stages will accept signals over a bandwidth many times this width, it is seen why this is so. For example, an amateur transmitter operating at 7 mc. may radiate a small amount of power at each of the harmonics (multiples) of this frequency. The amplitudes of these harmonics diminish rapidly with frequency, but multiples up to the sixteenth or eighteenth may be of sufficient strength to interfere with a weak television signal, depending upon the proximity of the amateur transmitter and its adjustment. Thus, with a harmonic falling every 7 mc., the transmitter adds a good chance of interfering with TV channels 2, 3, 4, and 5, since the 8th through 11th harmonics at 7 mc. fall within them. The degree of interference is usually determined by the proximity of the harmonic to the frequency of the picture carrier. If it is close, the harmonic must be weaker by about 50 db, to avoid interference.

By far the most serious harmonic interference is that caused by the secondary of primary transformers operating in the 28 mc. band, since this harmonic falls directly in channel 2 and is usually quite strong. Another such case of troublesome interference is found in the secondary of FM stations which fall within the high-band TV channels. The commercial solution to this problem has been similar to that adopted by amateurs — the use of filters to prevent spurious radiation.

TVI may also be caused by low-frequency signals getting into the receiver. If stages, either through the tuner or by direct pick-up in the audio circuit. Cases have been observed where picture reception was prevent-

**TVI Reduction at the Transmitter**

Of course, the most effective approach to interference elimination is to start at the source. The harmonic content of the transmitter signal is
temperately affected by circuit adjustment such as grid bias, grid drive, grid modulation percentage, and tank circuit L-C ratios. If the generation of harmonics and parasitics is first minimized by the selection of the proper values for these variables, the job of preventing the radiation of the remainder is considerably simplified.

In addition, it has usually proven necessary to completely shield the offending transmitter before the work of harmonic suppression by the use of filters can proceed. Otherwise, harmonic radiation may occur from the final tank coil and other parts of the transmitter. Since the wavelength of the harmonics which cause TVI are relatively short, leaks c or obsolete length may act as efficient antennas.

The need for shielding may be determined by loading the transmitter with a "dummy" lamp-load substituted for the antenna. If the TVI clears up, it indicates that the interfering signal is being radiated by the antenna and that the present degree of shielding is inadequate. If this test shows that more shielding is needed, the type required need not be elaborated, but must be complete. Commercially built metal cabinets, although neat in appearance, do not always provide effective shielding because the position of doors, cracks, and ventilating louvers are not always sufficient. The most popular method of shielding employed is amateur practice is to put a large transformer r.f. chassis in a box made up of closely woven metal screening, soldered at all junctions to make it absolutely r.f.-proof. This shielded chassis and panel may then be mounted in a standard rack or cabinet to improve the appearance.

With the r.f. portions of the transmitter completely shielded, it becomes a relatively simple matter to filter all leads entering this metal enclosure, as the manner indicated in Fig. 1. It must be remembered that the key or microphone feed is a potential source of r.f. leakage and must be either shielded or bypassed. Also, a.c. power leads which enter the chassis must also be filtered. For this purpose, a balanced single-pi-section, low-pass filter as shown in Fig. 2 may be employed. A unit of this type may be constructed in a small metal box and bonded solidly to the outside of transmitter shield box for maximum effectiveness. The line filter should not be accessible inside of the transmitter housing because of the danger of the components coupling to harmonics from the line circuit.

For d.c. leads which enter the shielded compartment, a simple low-pass filter of the type illustrated in Fig. 3 has proven effective in preventing r.f. leakage. The values of the components are not critical, but they should be of high quality. Inductors should be of a universal-wound type so that distributed capacitance is reduced. Mica capacitors should be chosen, according to voltage requirements. A filter of this kind should be used in each d.c. lead which might conduct r.f. out of the shielded housing. Like the line filter, these d.c. filters should also be assigned in a separate metal box which is fastened to the outside of the main shield compartment. A common housing may be used for all power and lead filters.

After the job of shielding the transmitter and filtering power leads has been completed, it should be checked again for TVI. If all signs of interference to nearby television receivers have diminished when full transmitter power is applied to a dummy load inside of the shielded compartment under conditions of full modulation or keying, this part of the job is satisfactory.

If the interference appears when the antenna is again connected, the TVI is reaching the receiver by radiation from the antenna. It may be of the harmonic type or the receiver overload type. At this point it is well to determine which, since further changes at the transmitter will not eliminate the latter type. The harmonic content of the transmitter signal may be checked by listening on the multivox of the operating frequency with a good VHF receiver, or by building a crystal "harmonic checker." The circuit of a simple device which will facilitate this is shown in Fig. 4. It consists of a parallel C-L circuit which tunes to the low TV frequencies and which is linked coupled to a crystal rectifier and indicating meter. The tuned circuit must be calibrated in frequency so that harmonics may be identified. Several of the commercial absorbors and monitors may be used as harmonic checkers by the addition of the crystal indicating circuit. Alternatively, a grid-dip meter of the type which has an attached grid-coupled oscillator tube as a grid detector may be employed for obtaining harmonics.

The harmonic checker should be loosely coupled to the output of the
transmitter and a systematic search for spurious frequencies made. The sensitivity of the indicator will be better if a low range microammeter is used. The frequency of all signals detected, other than the carrier, should be carefully tabulated, since this information will prove of value in determining filter requirements.

If any ratcheting is detected in the television bands, a filter between the transmitter and the antenna is necessary. Ideally, this filter should be a unit which transmits the amateur frequencies without loss, while presenting infinite attenuation to all TV bands frequencies. Actually, these conditions may be approached with modern low-pass filters of the "m-derived" type. With such networks of relatively simple design it is possible to obtain attenuations greater than 100 db, at all television frequencies. If high quality components are used, the "invasion loss" in the amateur bands below 30 mc. may be less than 3 db. In addition, the attenuation at any given frequency within the rejection band may be "peaked up" by special design. In this way, added attenuation may be provided at specific frequencies where harmonic output is greatest.

A practical low-pass filter for use with amateur transmitters is shown in Fig. 5. This network starts attenuating at 45 mc. and should provide over 60 db. attenuation at all frequencies above 55 mc. It consists of four sections: two series m-derived end sections, one constant-

**LOW-PASS TRANSMITTER FILTER**

**TVI Reduction at the Receiver**

A transmitter of another service may cause harmful interference even though its signal is in accordance with the best engineering practice. If it is located in the immediate vicinity of the TV set, as is usually the case with amateur transmitters, it may produce picture interference by overloading the TV receiver front-end and so produce local hums. This, of course, is not the fault of the transmitter, or of any amateur or commercial. It is merely a consequence of the close spacing between the transmitter and the receiver, and of certain deficiencies in the TV set front-end design. Most receivers now incorporate one or two high-pass filter sections between the antenna terminals and the first r.f. or mixer stage. This filter is intended to prevent the passage of strong, low-frequency signals into the tuner, but to accept the TV signals.

In many cases, where the TV antennas intercept a very strong low-frequency signal, the built-in high-pass filter may not provide adequate attenuation to prevent interference of the "overloading" type. It usually becomes the responsibility of the TV service technician to diagnose this trouble and to provide a cure. For this purpose several additional sections of high-pass filter may be necessary. Such filters are available commercially or may be made up. A typical design is detailed in Fig. 6. This filter is a balanced configuration for 300 om. "twisted-pair" and is of the double pi-section type. It is designed to have a high-pass cutoff at about 53 mc. so as to provide maximum attenuation to all amateur frequencies up to the ten-meter band (50 mc.).

The high-pass receiver filter should be unattached in a metal box similar to that described for the low-pass transmitter filter. Complete shielding is not too important in this case, since unshielded transmission line is used. High quality components should be selected. The capacitors should be of the mica variety. Two 10-microfarad units may be used in parallel to form the required capacitance value and to minimize load inductance. The coils are close-wound on a three-sixteens inch low-loss form and are center-tapped by twisting a half-inch loop in the center turn of each. Such loops are then connected out soldered to ground, leaving a quarter-inch lead. All coils should be shielded at least one inch apart to avoid coupling.

Receiver interference of the "i.f. channel" type mentioned above will be also reduced by the high-pass filter if it is being picked up by the antenna. However, interference of both this type and the "overloading" type may be reduced by a new ringer, or ringer stage, in the receiver. In such cases, a low-pass filter of the type shown in Fig. 3 should be used. It may also be necessary to improve the receiver shielding by adding a bottom plate to the chassis.
Television Reception at "Shadowed" Locations

One of the most prevalent problems confronting the rural TV viewer and his service technician is that of providing reception at remote sites well within the normal service range of one or more transmitters but "shadowed" by topographical details. In hilly or mountainous terrain, many communities find, indeed, whole cities, are in the shadowing position of having strong, steady television signals going by a few hundred feet overhead, but with little or no signal available to antennas of practical height in the valley. A typical situation of this kind is depicted in Fig. 1.

The social and economic implications of this common situation are many and seem a high price to pay for a wrinkle in the surface of the earth form-ing long before anyone had television transmission in mind. To the viewer, it means missing out on the educational and entertainment miracle of television. To the TV dealer and serviceman, it means wide fields of potential set sales and servicing jobs below. And, of course, to the television broadcaster and advertiser, it means reduced coverage. The following paragraphs are devoted to a discussion of some of the solutions which have or might be re-orted to in such instances.

There are several approaches which might be used in "fill-inimating" a television receiver or community situation as in Fig. 1. They include:

(a) A booster station located on the hill as "A" and relay-ing the signal on the same frequency.
(b) A satellite station situated at "A" rebroadcasting the signal on another frequency, such as a UHF channel.
(c) A "point-to-point" antenna at "A" receiving the television signal and re-radiating it into the valley.
(d) A community antenna located on the hill top with a transmission line distributor system feeding receivers in the valley.

In evaluating the applicability of these approaches to specific locations there are many legal, economical, and technical factors to be considered. We will now examine one of these.

The Booster Station

The operation of a relay transmitter by the distant TV station involves considerable legal and technical complication. Legally it would require a license by the Federal Communications Commission as well as an agreement with the television station whose programs were to be relayed. It also requires a source of electrical power at the relay station site as well as frequent or continuous attendance. In general, the expense involved makes this type of endeavor which must be financed on a commercial basis rather than run as a community enterprise. Nevertheless, experimental booster stations of this kind have been authorized by the F.C.C.

The block diagram of a typical equipment layout required by a booster station is given in Fig. 2. A high gain receiving antenna oriented to receive the signal of the desired TV station, to the exclusion of any other on the same channel, feeds a low-noise r.f. amplifier. This drives a linear power amplifier which builds the signal up to the level required for re-relay by a second antenna oriented to illuminate the desired coverage area. For single channel relay, the bandwidth of the overall system must be at most six megacycles. The total system gain will depend, of course, upon the signal strength available from the master station. In "fringe" areas at least 100 decibels of overall gain must be provided if the coverage angle is large. Usually the power fed to the transmitting antenna need only be a few watts.

One of the technical difficulties encountered in the operation of a booster re-relaying on the same channel as the master station is that of feed-back. Enough isolation must be provided between the output and input to prevent such regeneration. This is usually accomplished by nulling antennas with high cross-over-back rations placed back-to-back. Additional isolation is also available by plac- ing the receiving antenna and associated low-level pre-amplifier equipment a few hundred feet from the power amplifier and transmitting antenna. A high grade coaxial cable is used to interconnect the two.
One economic factor in favor of the booster approach is the fact that no expensive frequency control and sweep generator standards are required. Since the booster station is essentially a linear amplifier, all of the standards are established by the master station. Thus, the cost of an installation of this type is much less than that of a small station capable of originating programs, but is still prohibitive for the isolated viewer or small community.

Satellite Stations
Relaying the television signal on another channel frequency, such as a UHF channel, is even more complicated than some-channel relaying in many respects. In addition to requiring FCC authorization and master station permission, considerably more equipment is needed. Since the only manner in which the channel frequency can be changed is by heterodyning the master station carrier to a new frequency by modulating a locally generated carrier on the new frequency with the video signal of the master station, facilities for maintaining the required frequency stability will be required in either case. On the other hand, the problem of input-output isolation is eliminated in a retransmitting system of this kind since the frequency transmitted is different from that received from the master station. The satellite station can be arranged to select any one of several master station signals available and relay them on the allocated frequency.

Black diagrams of two possible equipments for satellite station relaying are shown in Fig. 2.

Passive TV Relaying
Another interesting possibility for television service where a strong signal from the master station is available might be called "passive" relaying. The essentials of this scheme are illustrated in Fig. 4. A high gain antenna situated on a high place within line-of-sight of both the transmitter and the shadowed receiver sites receives the signal from the TV station and feeds it to another high-gain antenna oriented to reradiate the signal into a relay. This system has proven practical in several instances. It has many advantages over the foregoing systems which put it within the reach of the single isolated receiving site or the small community.

Since this system is totally "passive," i.e., does not employ amplifying or transmitting equipment of any kind, the FCC approval has not been required. For the same reason, no source of electrical power is required at the relaying site and a minimum of maintenance is necessary. Those are decided advantages since the sites which are suitable for relaying of this kind are usually quite inaccessible.

Another advantage offered by the passive antenna system is that of multichannel operation. If several strong stations are located in the same direction, it will usually be possible to relay all of them simultaneously. In general, multiple-channel installations will require the use of more elaborate antenna arrays, however.

If the site available for relaying is insufficiently large the rubber horn antenna offers high gain and broadband operation while involving little expense. It can be constructed of wire, as contrasted to the more expensive aluminum tubing required in some other high gain designs. If the site is wooded, trees can usually be pressed into service to support the horns of the rhombics. Because the radiation pattern of a large rhombic is rather sharp, care must be taken to properly align both the receiving and the transmitting antennas. The major lobe of the rhombic is usually tilted a few degrees above the plane of the antenna. For this reason, it might be necessary to place the two antennas on opposite sides of the obstructing hill and interconnect them with a low-loss, high-impedance transmission line. Fig. 5 shows the dimensions of a rhombic antenna design which would be suitable at small locations. The gain of such an antenna increases with the number of wavelengths per leg.
Another antenna type which could be used is the "chicken-wire horn" illustrated in Fig. 6. This design provides all-channel operation and reasonably high gain while requiring less space than the rhombic. Two such antennas could be used back-to-back at the crest of the hill or separated by some distance and connected with 50 ohm line. Of course, the length of the line should be kept to a minimum since its losses detract from the gain provided by the antennas.

Other high gain TV antenna designs could be utilized for passive relaying. In instances where only single channel operation is desired, multi-element stacked Yagi designs, cut for the proper frequency will provide good gain and directivity in small space. Antennas of the "billboard" type should also be useful.

In the choice of the transmitting antenna, attention must be given to the angle of radiation. If the receivers are spread out over considerable area close to the re-relaying antenna, the pattern of this antenna must be wide enough to illuminate all of them. Since the object of the relaying system is to provide as much total system gain as possible, the antennas at the receiving sites must also be high gain designs of the variety usually employed in fringe areas. They, in turn, must be accurately oriented on the hill-top relaying antenna.

Community Antenna Systems

At locations where the receiving sites are only a few hundred yards from an elevated place where relatively strong television signals can be received, the community antenna scheme has been used successfully. As illustrated in Fig. 7, this system utilizes a high gain antenna on a high place and a long, low loss transmission line distribution system to "pipe" the signal to one or more receivers.

Technically, the problems associated with the community antenna approach are very similar to those involved in providing reception in large apartment buildings. The gain of the receiving antenna must be sufficient to offset the losses of the feed line, and provisions must be made to isolate the various receivers to prevent interaction between them. Cathode-follower isolation stages or resistive isolation pads are usually employed for the latter purpose. Special attention must also be given to lightning protection in such installations, since an antenna on a high place is especially prone to such phenomena.

Low losses, consistent with economy, is the main factor to be considered in the choice of the transmission line between the hill-top antenna and the receiver distribution point. Special consideration should be given to the high imbalance, open wire line and the surface wave transmission line. The cost of installation of the latter decreases with frequency, making it attractive for UHF use.
SECTION II
CIRCUITS

Electronic Oscillators
Part 1

In the past 25 years of progress, it is probably no other basic electronic circuit has increased in scope and versatility as much as the vacuum tube oscillator. Practically every radio receiver contains at least one oscillator, and TV receivers normally contain three. Every transmitter must contain a carrier-generating or frequency-controlling oscillator, and special types (such as those employing single-sided output) employ several in many cases. The wide use of oscillators in test equipment, such as audio and radio frequency signal generators, frequency meters, grid dip meters etc., and in magnetic recorders is well known.

The engineer, service technician, amateur and experimenter are thus vitally affected by the operation of oscillators in general, and important commonly used types in particular. The important fundamental concepts and design factors, and their application to every-day use of oscillators will be the objective of this discussion.

Definition of Oscillator

An oscillator is any device which can be induced into cyclic repetitive action. Mechanically on example is the clock pendulum; its electrical counterpart is a tuned resonant circuit. In both cases, the period of each cycle, and thus the frequency of oscillation is controlled, but energy must be added to overcome the loss in the device if sustained oscillations are to be obtained. Since we are primarily interested in sustained oscillations without damping, a complete oscillator must have two main parts: a frequency-controlling device which is usually a resonant circuit, and another part which applies energy to the frequency-controlling device in the proper manner to sustain oscillations. The latter is usually an amplifier.

The Institute of Radio Engineers defines an oscillator as:

A non-rotating device for producing alternating current, the output frequency of which is determined by the characteristic of the device. Standards on Automatic Modulation Systems and Transmitters, Definitions of Terms — IRE 1948.

This definition is broad enough to cover all electronic oscillators. We are concerned here with the electrical oscillator which is an electrical oscillator employing one or more vacuum tubes.

An electronic oscillator requires input energy to overcome tube and circuit losses and to supply the required output power. This input energy is obtained by means of electrical energy or from the plate power supply, and indirectly from the heater or filament current to the tube. Basically, it can be considered a converter more or less than a generator, since it connects electrical energy from one frequency to another usually higher than the input frequency.

Negative Resistance Requirement

For oscillation, energy must flow from the output (usually the plate) circuit to the input circuit (usually the grid) in such magnitude and phase as to overcome the losses of the system. But the basic amplifying action of a vacuum tube is to produce plate voltage which is approximately (exactly with a resistance load) 180 degrees out of phase with the grid voltage which produces it. Part of this output voltage must be applied to the grid circuit in phase with the grid voltage. This is done by inverting the phase (either actually or effectively) of that part of the plate voltage fed back to the grid circuit.

When this condition exists the network develops a negative resistance in the circuit. In a negative resistance, the current increases as the voltage decreases; thus the current and voltage changes are out of phase 180 degrees.

There are three main ways in which a negative resistance can be provided in vacuum tube circuits for oscillation:

1. By actually transmitting a desired portion of the output signal voltage to the input circuit in a feedback circuit which reverses the phase.
2. By the design of a tube, or the adjustment of the applied potential to the tube, so that it exhibits a negative resistance characteristic.

In this discussion, we will concern ourselves with the basic functional factors in types 1a and 2b. Type 2 will be considered later.

A. Inductively

B. Capacitively, through the grid-plate capacitance or external capacitance.

Criterion For Oscillation

In the consideration of any given oscillator circuit, it is important to know under what conditions of circuit design and adjustment oscillation will take place, as these conditions are limited.

For any oscillator of the feedback type, the Barkhausen criterion of Fig.
1 is applicable. This figure shows the oscillator broken up into its two main parts, the amplifier and the feedback link. The input voltage to the amplifier $E_{in}$ is the voltage fed back through the feedback circuit. This simple derivation shows that the fraction of the output voltage which is fed back ($B$) must be equal to the reciprocal of the gain. Both of these factors are complex, because both the amplifier and the feedback circuits do, in general, introduce phase shifts. For oscillation, the phase must shift an odd multiple of $2\pi$.

Equations 2 and 3 apply the criterion to a grounded-cathode amplifier. This expression is general, and can be applied to any particular circuit by evaluating $\frac{B}{1+G(B)}$ in terms of the circuit parameters involved and substituting them in the general expression equation 3.

### Industrial Feedback Circuits

From a theoretical standpoint, probably the most direct method of providing negative resistance for oscillation is by mutual inductance between coils in the plate and grid circuits respectively. The two most common circuits of this type are illustrated, along with their equivalent circuits, in Fig. 2 and 5 respectively.

### Tuned Grid Oscillator

$$\frac{w_e}{\sqrt{1+A}} = \frac{1}{\sqrt{L_p C + 1}} \quad \text{Eq. 4}$$

$$M\frac{\omega}{\sqrt{1+1}} = \frac{L_p}{C} \quad \text{Eq. 5}$$

$$\frac{M}{A} = \frac{1}{1+A} - \frac{L_p}{C} \quad \text{Eq. 6}$$

Where $A = \frac{L_p R_g}{L_g}$ and $\frac{M}{A} = \text{resonant freq.}$

### Tuned Plate Oscillator

$$\omega = \frac{\sqrt{L_p}}{R_g} \quad \text{Eq. 7}$$

$$M = \frac{L_p}{C} \quad \text{Eq. 8}$$

$$\frac{M}{A} = \frac{L_p}{C} \quad \text{Eq. 9}$$

These frequency equations are useful primarily in a qualitative way; quantitatively since $w$ is ordinarily very close to $\omega_0$, the values of $L$ and $C$ are usually adjusted at least partly.
by empirical means, starting with values which by themselves resonate at the frequency of oscillation. But these expressions are important in indicating the direction of frequency change with change of $Q$ and external loading.

In comparing the two circuits, the expression $A$ is significant. Because the feedback "higher"-grid to the tuned grid circuit is usually much smaller than $Q$ and $Q$ is very much smaller than $V$, the value of $A$ is far less than 1. Because of the presence of the ratio of inductance in $A$, and the fact that the plate coil is normally much smaller than the grid coil in the tuned grid circuit, it will be noted that the frequency of this circuit is less sensitive to changes in $R$ (and thus $Q$) than in the tuned plate circuit. Also, in the expressions for $M$ and $C$, the plate inductance appears in the numerator of the first fraction for both oscillators. Accordingly, since the plate inductance is relatively much smaller in the tuned grid circuit, the latter will oscillate with smaller values of $M$ and $C$ than will the tuned plate circuit.

In addition to the above-mentioned relative advantages, the tuned plate circuit requires of its designer that he make the unpleasant choice between (1) having plate 6-volt applied to grid and an r-f choke with series feed or (2) adding an r-f choke, with its added expense and danger of self-resonance somewhere in the tuning range, with shunt feed. On the other hand, in the tuned grid circuit, the plate coil is apologetic, isolated from tuning adjustments, is easily insulated and adapts itself nicely to series feed, which is always used.

In defense of the tuned plate circuit, it should be said that it is less sensitive to power supply voltage variations. This arises from the fact that space-charge capacitance, a function of plate voltage, is greater between grid and cathode than between plate and cathode. The space-charge effect is thus greater upon the frequency where the frequency-determining circuit is connected to the grid than when it is connected to the plate.

It is important to note some of the assumptions made in the derivation of the equations 4 through 6. First, the effect of resistive grid current, present in nearly all oscillators, has been neglected. The vacuum tube and its circuit has been considered as a linear device, whereas ordinarily it must be non-linear for oscillator operation. However, this is not too bad an assumption. Nowadays, the power oscillator is a thing of the past except in special applications.

and the usual circuit is designed for stability and flexibility. For the attainment of the best stability, the grid current must be kept relatively low, making the equations nearly valid.

Another assumption in the analysis of the inductive feedback oscillators is that grid-plate capacitance is negligible. This is a reasonable assumption, since, although it does add a certain amount of loading effect to the input circuit, this capacitance does not materially affect the action of the inductively-coupled feedback.

In general, in advance of inductively-coupled feedback oscillators is that $M$ provides a convenient parameter for adjustment of operating characteristics by adjustment of the size of the feedback coil and its physical position. A general disadvantage in multi-range circuits is that bias switching is complicated by the additional coil terminals.

Capacitor Feedback Circuits

Under certain conditions, a deliberate circuit feedback path is not necessary for the support of oscillations. One common instance of this is the regenerative effect, especially in triodes, of the grid-plate capacitance. This effect becomes evident upon examination of the expression for the input resistance of a triode.

Figure 4 illustrates the input impedance, which includes in general, both a resistive and a capacitive component. The values of these depend upon the nature and magnitude of the plate load impedance as well as the grid-plate capacitance. When the plate load is a pure resistance, the capacitive component of input impedance becomes infinite and the input impedance becomes a pure capacitance. (Amplification $K$ depending upon plate load impedance.)

Under certain conditions, a deliberate circuit feedback path is not necessary for the support of oscillations. One common instance of this is the regenerative effect, especially in triodes, of the grid-plate capacitance. This effect becomes evident upon examination of the expression for the input resistance of a triode.
However, when the plate load impedance becomes inductive, and $K_1$ becomes negative, the input resistance becomes negative, if the negative resistance exceeds grid losses, oscillation can take place. Thus a simple amplifier can become an oscillator if the plate load is inductive and the grid-plate capacitance is sufficient. As can be seen from Eq. 10, the frequency, grid-plate capacitance, and phase shift of gain are all interested in determining whether the input resistance is to be negative and oscillations will take place.

As in our previous discussions of inductive feedback oscillators, the effect of grid current is neglected and the tube is assumed to be a linear device, both permissible for most practical oscillators. It is also assumed that grid-plate capacitance has negligible effect on the gain.

About the only common type of oscillator depending primarily upon grid-plate capacitance for oscillation is the tuned-plate tuned-grid type illustrated in Fig. 5, with its equivalent circuit. In essence, it is simply a tuned-circuit amplifier adjusted to oscillate. Sometimes an external capacitor is connected between plate and grid; its purpose would be either to increase feedback at low frequencies or to improve stability by reducing the effect of variations in the grid-plate capacitance.

From the equivalent circuit it will be noted that the feedback coupling is the result of the fact the grid circuit and the grid-plate capacitance are connected in series across the plate signal voltage.

As was explained above, this type of circuit will oscillate if the plate load is inductive. Since the plate load here is a parallel tuned circuit, the resonant frequency of this tuned circuit must be made higher than the expected frequency of oscillation. The net reactance of the parallel combination is then inductive as desired.

It can also be shown that the grid tuned circuit must be tuned to be inductive, but slightly less inductive than the plate circuit.

The basic effective setup can perhaps be more clearly visualized by substituting inductances of the effective value for the two resonant circuits respectively, as illustrated in Fig. 6. It can be seen that for steady-state oscillation conditions, the effective inductances of the two tuned circuits and the grid-plate capacitance must resonate at the oscillation frequency.

The tuned-plate tuned-grid oscillator has the disadvantage of depending upon the grid-plate capacitance of the tube for a vital part of its operation. Its stability of frequency is thus affected by the thermal and other causes of variation of this factor. Besides, the fact that the grid-plate capacitance is fixed causes the degree of feedback to vary in an undesirable manner when an appreciable frequency range is to be covered. These difficulties, added to the inconvenience and expense of providing two tuned circuits, are undoubtedly the reasons that this circuit is not often encountered.

A slight variation of the tuned-plate tuned-grid circuit is the "TNT" version, in which the principle is the same, but the grid circuit is adjusted to its proper effective inductance by the distributed capacitance of the grid coil, instead of the grid capacitance. Although this eliminates the need for one capacitor, this circuit introduces all the other disadvantages of the tuned-plate tuned-grid circuit.

**Tuned Circuit Feedback**

Other types of oscillators do not employ either inductive or capacitive feedback in the manner described above, but derive the feedback phase and amplitude relations from a tuned circuit. This tuned circuit is ordinarily a single one which determines the frequency of oscillation.

Probably the best known example of this type is the Hartley, illustrated with its equivalent circuit in Fig. 7. The tuned circuit is divided into two parts by the cathode tap. Grid and plate signal voltages of opposite phase are then obtained from the two ends, respectively, of the resonant circuit. By analysis of the equivalent circuit in the same manner as for the inductive-feedback circuit earlier in this article, the following relations are obtained:

**Figure 7**

HARTLEY OSCILLATOR
HARTLEY OSCILLATOR

$$\omega = \frac{1 + \frac{E_i}{R_p}}{\sqrt{(L_1 + L_2 + 2N)C}}$$

[Eq 12]

$$gm = \frac{C(R_1 + R_2)(L_1 + L_2 + 2M)}{(L_1 + L_2)(L_1 + M)}$$

[Eq 13]

$$if \ M = 0: \ gm = \frac{C(R_1 + R_2)}{L_1}$$

[Eq 14]

where \( R = \) total tuned circuit resistance (in \( L \))

\( L = \) total tuned circuit

$$M = \frac{L_1 + M}{L_2 + M} + \frac{C}{R_1 R(L_1 + L_2 + 2M)(L_1 + L_2)}$$

[Eq 15]

Note that the frequency relationship Eq. 12 is the same as that for the tuned plate oscillator (Eq. 7) except that in this case it is the resistance of the plate section of the coil, instead of the total resistance of the coil which influences frequency of operation. Thus it would be expected that for equivalent coils, the stability would be better for the Hartley.

From the expression for \( u \), it can be seen that the greater the ratio of the grid-to-cathode section of the coil to the plate-to-cathode section, the easier it is for oscillation to take place. However, the plate section cannot be too small, since then there will not be sufficient transfer of energy from the plate to the cell.

It is interesting to note that it is not necessary to have either grid-plate capacitance or mutual inductance between the sections of the coil to support oscillation. The expressions are derived containing \( M \) because such \( M \) is usually present. However, if the two sections of the coil are entirely separated oscillation takes place. In fact, oscillation is even more vigorous without mutual inductance. Of course, the number of turns in each section of the cell must be somewhat greater, to make up for the loss of \( M \), if the same frequency is to be maintained with the same tuning capacitor.

The Colpitts oscillator is a variation of the Hartley principle in which the tuned circuit is divided by a capacitor voltage divider instead of a tap on the coil. It is shown with equivalent circuit in Fig. 8. The expressions for this oscillator are as follows:

$$w = \frac{1}{\sqrt{l_1 R_p + C_{1} + C_2}}$$

[Eq 17]

$$\mu = \frac{C_2}{C_1} + \frac{1}{\pi R(C_1 + C_2)}$$

[Eq 18]

$$gm = \frac{R(C_1 + C_2)}{L}$$

[Eq 19]

Note that the tuned circuit must be adjusted for a resonant frequency slightly below the actual frequency of oscillation. The expressions for \( w \) and \( R \) are similar to those for the Hartley, except that they contain the divider capacitances instead of the sections of the coil. The values of both \( s \) and \( a \) necessary for oscillation are small. In the expression for frequency, the capacitance values play an important part. It is noteworthy that a relatively high value for \( C_2 \) makes for less easy oscillation and poorer stability.

One of the important advantages of the Colpitts oscillator is the relatively large capacitances (\( C_1 \) and \( C_2 \)) are shared across the plate-to-cathode and grid-to-cathode interelectrode capacitances of the tube. This mini-

**FIGURE 8**
Electronic Oscillators

Part 2: Local Oscillators in A-M Receivers

A great majority of the local oscillators in AM receivers are part of converter-tube circuits. In short-wave communications receivers, separate-tube arrangements are frequently encountered, especially when the coverage extends to frequencies of 30 mc or higher.

Arrangements Used.

The most popular oscillator circuit in AM receivers is the Hartley, although it does not usually appear in its basic form. Most frequently it is found as the grounded-plate version.

A typical pentagrid converter oscillator circuit is shown in Fig. 1 (A). The basic triode grounded-plate Hartley is shown at (B) for comparison. In the pentagrid converter, grids 2 and 4 take the place of the triode plate.

The circuit variation of Fig. 2 is particularly popular in small, low-priced AM receivers. An additional winding L1 is interwound with L2. L1 is called a “bifilar” winding. One end is left open. The capacitances between L1 and L2 takes the place of the capacitor C1 in Fig. 1.

Two examples of the use of inductive feedback in AM receivers are shown in Fig. 3 and 4, respectively. The circuit of Fig. 3 employs the pentagrid converter version of the “tuned grid feedback” type. Figure 4 shows a cathode-coupled feedback arrangement. The latter can be pictured as the circuit of Fig. 3 with the plate feedback coil moved through the plate power supply and into the cathode circuit.

Inductive feedback circuits are not very frequently encountered in medium and higher-frequency receivers because of the expense of the additional coil winding and its connections, and because of the inconvenience of switching frequency bands. However, at low and very low frequencies, inductive feedback is often employed because of the difficulty in obtaining sufficient feedback other ways.

A typical dual triode oscillator-mixer circuit is shown in Fig. 5. This arrangement is popular in TV and communications receivers where low noise level and minimum oscillator-RF interaction are required. Sometimes a small capacitor connected between grid or cathode of the oscillator and mixer grid is used for injecting oscillator voltage to the mixer. In other cases, the cathode of the mixer and oscillator are either common or coupled together.

Requirements.

The following are important requirements in the design of oscillators in low-frequency AM receivers:

1. Ease of oscillator
2. Freedom from undesired resonances
3. Constant output amplitude
4. Frequency stability
5. Minimum of harmonic output
6. Tracking

Ease of Oscillation

Conditions for oscillation must be well fulfilled, so there is no tendency toward delay or failure in starting or maintaining oscillation. The nature of these conditions was discussed in the first article of this series. Oscillation criteria show that from a general theoretical standpoint, the tuned grid feedback and Hartley are the easiest oscillators. The Colpitts circuit also oscillates easily, but is seldom used in receivers because of

![Figure 1: Versions of Grounded-Plate Hartley Circuit](image_url)
the added components required (capacitors in the tank circuit).

If the basic design does not allow easy oscillation, then excessive plate and/or grid currents may be neces-
sary, with resulting overheating and instability as well as excessive power requirement.

Undesired Resonances
Undesired resonances are most likely to occur in the inductive feedback type of circuits (tuned grid and tuned plate types). In these circuits the feedback or tickler coil may reso-
nate with its distributed capacitance or with stray circuit capacitances. If the resonant frequency is within the tuning range, sufficient power may be absorbed to stop oscillation at and around that frequency. At least op-
eration in the vicinity of the fre-
quency of undesired resonance be-
comes unstable and undependable.

If the receiver is of the multi-range type, the coil of one range which is unused may self-resonate within the tuning range of the coil in use, with results similar to those mentioned above. The unwanted frequencies of resonance and their effects are often referred to as "dead spots" and "null-outs".

Undesired resonances can, to an appreciable extent, be avoided by careful initial design. However, all possible resonances naturally cannot be anticipated, so a thorough test for null-outs is a sensible precau-
tion. The best test is observation of the value of rectified grid current as the oscillator is tuned through its range. The rectified current of the mixer injection grid is also a good indicator if separate oscillator and mixer tubes are used. Any tendency toward unwanted resonances will show up as sharp variations of this grid current as the resonant frequen-
cies are approached. Such condi-
tions can also be traced with a grid-
dip oscillator, but it must be remem-
bered that such an analysis is not complete unless the tuning capacitor of the tested oscillator is varied through its complete range. Some-
times the tuning capacitor is part of the undesired resonant circuit.

No general formula for eliminating unwanted resonances can be given; it's just a matter of changing the circuit constants so that these reson-
ances are moved outside the tuning range, or better, but seldom possible, eliminated altogether. In multi-
range receivers in which 2 or more coils of different ranges interfere with each other, the resonant frequency can be
moved out of the range by adding a section of the switch which shortens each tuned coil. On some ranges it may be better to leave the coil open when unused. In any event, a large percentage of troubles can be avoided by careful initial study of the inductances and capacitances involved, and the checking of each coil as to its self-resonance and its mutual inductance with other coils after as-

stallation in the circuit.

Constant Amplitude
The amplitude of oscillator injec-
tion voltage has an important effect
on the operation of a superhetro-
dyne receiver. If the amplitude is ex-
ncessive, and the mixer is driven be-
yond cutoff, sharp discontinuities in the conversion characteristic occur, and excessive oscillator harmonics re-
sult. These harmonics lead to many spurious responses, manifested by whistles and "biting" in reception.

On the other hand, if the injection voltage is too low, conversion trans-
conductance falls off sharply, and re-
ceiver sensitivity is limited.

Thus it is important that the out-
put amplitude of the oscillator re-
mains constant over the tuning range. Unless compensated, the output of a capacitance-tuned oscillator increases as it is tuned from the low to the high frequency end of its tuning range. To compensate for this, some method of reducing relative output toward the high frequency end must be em-
ploved.

One convenient method employs the grid-cathode capacitance of the tube, with an added resistor to form a voltage divider, as shown in Fig. 6. Since the grid cathode capacitance portion of the divider is a lower im-


The circuit of Fig. 6 also compen-
sates amplitude by its loading effect on the grid coil. The resistor and capacitor in series damp the coil with a shunt resistance which decreases
with frequency, thus compensating for the normal oscillator amplitude increase. The effective shunting resistance is the reciprocal of the real part of the admittance of the two components in series, which can be derived as

$$R_{eff} = \frac{1 + \omega^2 C Q R^2}{\omega^2 C Q R^2}$$

For example, if the tube has a grid-cathode resistance of 10 ohms, and 2000 ohm resistor produces the following loading effect when connected in series with it:

- at 500 kHz: 840,000 ohms
- at 1000 kHz: 755,000 ohms
- at 1400 kHz: 115,300 ohms
- at 10 MHz: 10,100 ohms
- at 100 MHz: 252,000 ohms

Obviously, the value chosen for the compensating element depends upon the loading effect Q of the circuit as well as the degree of original amplitude variation to be compensated.

Other methods for amplitude stability have been employed, but in most AM receivers, the variations of the audio signals are prevented by use of any elaborate arrangements. Often the cartridge of the medium is included so that to vary the minimum allowable for sensitivity at the low frequency end and to maximum which will sensibly limit spurious response at the high end.

**Frequency-Stability**

Since the frequency of the IF signal, and thus proper reception, depends upon the frequency of the oscillator signal, frequency stability is very important. Its importance increases as receiver selectivity relative to received signal bandwidth increases.

For example, suppose an AM broadcast station transmitting audio modulation signals up to 12 kc is to be received. If the receiver has a pass band of just 24 kc, any drift in the local oscillator causes attenuation in the receiver of some of the high frequency modulation signal components. But suppose the same receiver is employed for a communications signal in which modulation frequencies are limited to a maximum of 3 kc, at a bandwidth of 6 kc. If this signal is originally tuned exactly in the center of the 7.2 kc receiver pass band, the oscillator can drift 3 kc either way without sideband clipping.

Oscillator drift is thus not as annoying if extra pass band is available at the receiver. However, in practical design, such extra response width is not ordinarily feasible because of resulting increased noise and because of adjacent channel interference effects.

Accordingly, appreciable attention must always be given to frequency stability. There are many factors involved in stability and probably more work has been done with it than with any other feature of oscillator operation. Since frequency stability is discussed in detail in a later article covering receiver oscillators, we shall not attempt full analysis at this point.

Frequency instability may arise from one or more of the following causes:

1. Variation in tuned circuit resonant frequency due to changes in L or C resulting from changes of temperature or humidity.
2. Changes in interelectrode capacitance values in the tube.
3. Fluxtuation of operating voltages.
4. Changes in the load offered by the mixer.

One instability effect important in connection with oscillators in AM receivers is known as "piling." Efforts to align the 7-tuned circuits to a received (or signal generator) signal result in a shift of oscillator frequency, detuning the receiver from the desired signal. This makes it difficult to obtain optimum alignment.

The interaction is due to (1) wrong coupling in the converter tube and (2) any direct coupling present between the oscillator and signal grid coils or circuits.

If "piling" cannot be completely eliminated in design, its effect can be
minimized by the "rocking" method of alignment. Better still, if a noise source is available, alignment by noise output will be free of pulling effects.

Harmonics

Harmonic output from the local oscillator is undesirable because it can lead to a number of different interference effects. Chief among these is the heating of an oscillator harmonic against a higher frequency station from which the harmonic is segregated by the intermediate frequency. Harmonics also adversely affect stability of the fundamental.

A certain amount of harmonic output is inevitable for all rf oscillators since non-linearity is essential to their operation. However, harmonics can be kept very low by careful attention to the following factors:

1. Use of a low L/C ratio. The selectivity of the tuned circuit is then much greater because of the high Q of the tuned circuit. Unfortunately, in capacitance loaded oscillators, a low L/C ratio at the high frequency end of the tuning range is not consistent with full tuning range possible from a given tuning capacitor. Ordinary tuning capacitors, when connected into an oscillator circuit, provide a maximum tuning ratio of about 3 to 1. This ratio is attained only at the extremity of a relatively high L/C ratio at the high frequency end of the tuning range. This problem can be overcome by inductance tuning, but such tuning is not practical or desirable in many receiver applications. Thus if harmonic output is to be minimized in wide-range tuning, other harmonic-reduction measures must be considered.

2. Increased coil Q. Higher effective Q during operation can of course be obtained by increasing coil Q as much as possible. Use of Litz wire within its favorite frequency range, use of bauq and other special windings, and optimum dimensional relations are well known methods. In the appropriate frequency ranges, addition of a powdered iron core can provide an appreciable increase in Q.

3. Limited power and drive. In most AM receivers, injection power requirements of the mixer are low enough to allow good operation with relatively low oscillator power output. Since harmonic content is greatly increased by use of large bias and drive voltage, it is desirable that feedback be reduced to a minimum necessary for easy oscillation. For lowest harmonic content with acceptable output, most oscillators should approach class B operation. The design procedure would be as follows:

(a) With no limitation on feedback path, adjust operating conditions, including feedback ratio, for maximum grid current. This would include adjustment of the tap in Hartley coil, capacitors in Colpitts divider, etc. The objective of this step is to obtain optimum feedback phase relation.

(b) Reduce applied plate and/or screen voltage to minimum value necessary for easy oscillation, also simultaneously decouple the resonant circuit to minimize grid current for desired output. Greatest reduction of feedback amplitude with maintenance of optimum feedback phase should be the objective.

(c) The frequency range of the oscillator is such that a suitable sensitive radio receiver covering the fundamental and several harmonic frequencies is available, the method of Fig. 8 can be used for testing harmonic output. The signal generator should be of the laboratory type, with low leakage and dependable attenuator and output voltage readings. The coupling to the tested oscillator should be light and through a shielded link with as little passive coupling as possible. The receiver must be well shielded, and should have a signal meter. If it
Electronic Oscillators

Part 3 - Tracking

The local oscillator frequency must, of course, come so as to be al-
ways separated from the resonant fre-
quency of the r-f tuned circuits by an amount equal to the intermediate 
frequency. This means that the oscil-
lator tuning range in kilocycles or megacycles must be exactly equal to the 
tuning range of the r-f tuned cir-
cuit.

Suppose the r-f circuit and the os-
cillator circuit are both tuned by 
identical sections of a gang capa-
tor, and the oscillator coil induc-
tance made smaller accordingly. The 
ocillator frequency cannot be kept 
equally spaced from the resonant fre-
quency of the r-f section without spec-
ial tracking measures. Instead, the oscil-
lator frequency varies as indi-
cated by curve b of Fig. 1 where 
curve a is that of the r-f tuned cir-
cuit. Note that the actual heterodyne 
signal frequency changes from a rela-
tively small difference frequency at 
the low end of the range to a rela-
tively high difference frequency at 
the high end. It is the purpose of 
tracking circuits to make the oscil-
lator frequency variation such that 
the r-f to oscillator frequency differ-
ence approaches a constant value.
equal to the intermediate frequency, as illustrated by curve c.

One simple way of providing tracking is by use of a smaller gang capacitor section for the oscillator, and "shaping" the rotor plates of the section. This shaping changes the capacitance vs rotation relation of the oscillator section of the capacitor so that it compensates for the deviation of curve b from the desired curve c. A typical shaped plate appearance is illustrated by Fig. 2. This method of tracking is widely used in low-priced AM broadcast receivers. It has two limitations: (1) it cannot be used in receivers providing more than one frequency band, because the plate shaping is good for only one range and (2) it provides no alignment adjustment for future correction for changes in the circuit constants due to temperature, humidity, dust, aging, etc.

Because of these limitations, the best available tracking method is considered to be that in which two adjustable capacitors are added to the oscillator circuit, as illustrated in Fig. 3. One is connected across the variable tuning capacitor (Cv) and is called a "shunt trimmer" or sometimes just "trimmer." The other is connected in series with one of the leads between the variable tuning capacitor and the oscillator coil (Cr) and is called a "series pad" or just "padder."

At the high frequency end of the tuning range, the tuning capacitor C is at its minimum value, and the shunt trimmer has a relatively appreciable capacitance. Accordingly, the shunt trimmer has marked effect on the oscillator tuning curve at and near the high frequency end of the range, but negligible effect toward the low frequency end, where the tuning capacitor value becomes very large in comparison.

On the other hand, the series pad has an important effect at and near the low frequency end of the tuning range and negligible effect toward the high frequency end. This follows from the basic fact that when two capacitors are connected in series, variation of one capacitor has its greatest effect on the combined series capacitance when the other capacitor is at its maximum value.

Thus, when the oscillator tuning inductance is properly chosen, the shunt trimmer corrects the high frequency portion of the tuning curve, and the series pad corrects the low portion. If the tracking is required, both of these adjustments are necessary to overcome production tolerances in inductance, distributed capacitance and stray capacitance.

Method of Calculating Tracking Component Values

In order to ensure good tracking, the designer must compute the proper values of L and Cr to make the tuning capacitor tuning curve approach perfect relation to the rf tuning curve. To do this, expressions can be set up for the resonant frequency of the combination of the three components in the tuned circuit plus the tuning capacitor section for each point at which oscillator tracking error is to be zero. Absolutely zero tracking error is of course not possible over the entire tuning range. However, the maximum error anywhere can be kept small by calculating circuit constants to give zero error at one or more frequencies. If these frequencies have been properly chosen, the remainder of the tuning curve will be very close to that desired.

The quality of a tracking circuit can be expressed by plotting the oscillator frequency error against frequencies in the rf tuning range of the receiver, as shown in Fig. 4. Positive errors (oscillator frequency too high) are plotted above the line, negative errors (oscillator frequency too low) below the line. The graph of Fig. 4 shows approximately the error variation resulting from the untracked curve of Fig. 1, curve b.

It can be shown that, as L, Cr and Cr approach their proper values, the error curve takes the approximate form shown in Fig. 5. Because of the three variables, the curve equation is a cubic, and crosses the zero axis in three places. These places are the frequencies of zero error. The proper locations of these frequencies can be determined by making certain assumptions. First, it is assumed that tracking is nearest perfect when the maximum error is the same at all three maximum points; in other words when the curve is symmetrical. By substituting boundary conditions in the general cubic equation, one derives the fact that the zero-error frequencies should be...
where $f_1 = f_2 = (f_c - f_0)$

$\Delta f = f_2 - f_1$

$f_c = f_1 + (f_c - f_0)$

where $f_1$, $f_2$ and $f_0$ are the pre-error

$f_c$ is the center frequency of the r-f

$tuning$ range

$f_0$ is the frequency at the low end

of the r-f tuning range

For the standard U.S. A.M. broad-

cast band of 550-1600 kc, these fre-

quencies come out to be $f_1 = 650.35$

kc, $f_2 = 1075$ kc, and $f_0 = 1520.65$

kc. Of course, individual manufac-

turers may have other considerations

which have led to the use of different

tracking frequencies; 800 kc and 1400

kc are widely used for the low and

high points. These points $(f_1$ and

$f_2$) are the ones at which the pad-

der and trimmer capacitors, respec-

tively, are adjusted. The center

point should then fall automatically

into place, providing the inductance

is correct.

In the derivation of design expres-

sions for the values of the shunt and

series padder and the oscillator coil

inductance, a general expression for

the resonant frequency of the oscil-

lator tank including the padder and

trimmer plus tuning capacitor is set

up. This general expression is then

modified to form three expressions,

one for each tracking frequency.

These are the center of the r-f tuning

capacitor value usually the same as

the oscillator cavity frequency of tun-

ning inductance. Solution of these

three expressions simultaneously re-

sults in equations for the trimmer,

padder and inductance. These mathe-

matical operations and the resulting

expressions for trimmer and padde-

r capacitance and inductance are quite

numerous, so will not be repeat-

ed here. However, several approach-

es will be found in the literature.

Practical Modifying Factors

The previously-described treat-

ment is one of several similar ap-

proaches to the problem. As men-

tioned above, other positions may

be assumed for the end tracking fre-

quencies, so they are within the tun-

ing range instead of at its edges. Thus,

it is common practice to use

600 kc and 1400 kc as low and high

frequency alignment points, respec-

tively, for the 550-1600 kc A.M broad-

cast band. The designer may also call

for the center "anti-error" point to fall

at the geometric mean (895 kc for

the above range) instead of at the

arithmetic mean frequency. Of course,

the center tracking frequen-

cy is important only in initial design.

If the latter is correct the zero

point automatically falls in place

in subsequent alignment.

In practice, there must always be

a certain amount of "cut-and-dry" ad-

justment of design values, after the

latter have been theoretically deter-

mined. This is necessary because of

modifying factors such as mutual in-

ductance.

One of the assumptions of most

analyses of tracking is that both the

signal and oscillator tuned circuits

are isolated and that the tuned coil

and connected capacitance are the

sole factors determining resonant

frequency. Without this assumption,

the mathematically derived expres-

sions would become overly cumbersome.

Obviously, signal and oscillator
coil would be of no use if not con-

nected to anything. The practical ap-

proach is to minimize the reactive

component due to coupling, so the

calculated tracking values still have

practical meaning.

In these circuits, this would seem
to favor the use of inductive coupling over capacitive coupling.

It would also encourage the use of an electromagnetic shield between pr-

imary and secondary windings which tuned r-f transformer. By the same

token, the oscillator would preferably be inductively coupled to the

mixer, where separate inductors are employed, and such coupling would be

as loose as possible. In pentagrid

converters, a certain amount of

space-charge capacitance coupling is of course inevitable.

When the oscillator is of the in-
ductive feedback type, serious modi-

fication of the tracking design can

result if the tickler coil has too much

inductance. Tight coupling with low

inductance is best, so that the self-

resonant frequency of the tickler

coil is well above the tuning range

limits. If the coupled reactance is

too large, it also limits the tuning

range, an important factor when the

latter is required to be relatively

large.

Use of Short-tuned Coils

In the idealized three-point track-

ing arrangement of Fig. 5, it is as-

sumed that, although the series and

shunt capacitance is adjustable dur-

ing alignment, the inductance is fix-

ed. This is true in a majority of

cases.

Thus, in practical design, the en-

gineer may anticipate what toler-

ances in fixed inductance value can be

allowed consistent with permis-

sible tracking error. Figure 6 shows

the effect on three-point track-

ing of changes in inductance.
assuming that, each time the induc-
tance is changed, the trimmer and
padlock capacitors are re-adjusted
in the predetermined tracks or tracking
frequencies. Note that as the induc-
tance gets far away from its proper
value, there becomes only one point
of maximum tracking error, instead of
the ideal case. The new maximun-error
point is also shifted in frequency from
its proper value

In some receivers, notably those of
the communications type, the oscil-
lator coil is " slug tuned " to its induc-
tance can be varied for alignment

The last decade has seen a vast
development of the frequencies above
20 mc., particularly the VHF
(30-300 mc) and UHF (300-3,000 mc)
ranges. The most important influ-
ces in this development have been
to radars, aircraft communications,
and FM and TV broadcast services.

In all these, oscillators play a vital role.
This Part discusses the features of oscil-
lators designed for operation in these
ranges and using vacuum tubes of con-
ventional design.

Special Problems at Higher
Frequencies

Vacuum tubes of conventional
(although sometimes somewhat mod-
ified) design are now being used in
commercially-available equipment as
oscillators operating as high as 1,000
mc. and above. However, successful
operation in the VHF and UHF re-
gions of the spectrum requires that
these difficulties be overcome.

These difficulties arise from vacuum
tube factors, and circuit factors, which,
although not noticeable as
low frequencies, take on special
importance as the frequency
ranges. Figure 1 illustrates these
factors, which, are as far as the

tube itself is concerned, transit time,
lead inductance, and interelectrode

capacitance. These electrical tube
factors are discussed first, followed
by the additional important circuit and
physical factors.

Transit time is the time it takes
an electron in the tube's electron
stream to travel from cathode to plate.
If this transit time is appreci-
able compared to the period of 1

cycle at the desired frequency of
oscillation, it is extremely difficult
to sustain oscillation. This is be-
cause, as the transit time approaches
the period of 1 cycle, the phasing
between plate and grid voltages is
affected in such a way as to intro-
duce the effect of shunting resistance
(conductance) between grid and cathode.
Since all or part of the tuned circuit is connected or
coupled between grid and cathode, the oscil-
lating circuit is adversely loaded by this
resistance effect. An undue
amount of power may thus be dis-
sipated, and in severe cases (higher
frequencies and unwattable tubes)
sufficient energy cannot be fed back
to sustain oscillation. Many tubes
which have input impedances as high
as 2 to 20 megohms at low
frequencies (below 2 mc) have val-
tes as low as 20 to 200 ohms at 300
mc and higher.

Transit time is obviously a func-
tion of the spacing between the cation-
and the plate of the tube; the
greater the spacing, the longer it
takes for the electrons to traverse
the gap. It is also a function
of the relative grid-to-cathode spacing,
which is the harmful N.B.E.;
resulting from transit time, is di-
rectly proportional to Cn and in-
versely proportional to the square
of the frequency. However, the Cn
must be kept high to support oscil-
lation and provide stability so the
transit time must be kept low by
minimizing spacing and interelec-
trode capacitance. An increase in
plate voltage reduces transit time
by speeding up the electron stream,
but increasing plate voltage over its
rated value is likely to overload the
tube, and so is not a satisfactory
method.

The magnitude of the effect of
transit time on input loading can be
gauged from the following ex-
pression:

Electronic Oscillators
Part 4: VHF and UHF Oscillator Circuits

The magnitude of the effect of
transit time on input loading can be
gauged from the following ex-
pression:
where: 
\[ E_{i} = K \phi \times f^{2} \]

- \( E_{i} \) = grid input voltage due to transit time.
- \( K \) = tube transconductance.
- \( \phi \) = frequency of oscillation.
- \( f \) = transit time from cathode to grid.
- \( \epsilon \) = conductance of the tube.

Although this expression is derived for a normal grid, it is just as useful qualitatively in the case of an oscillator.

Note that the input conductance increases (resistance decreases) with the square of the frequency. Thus the input resistance of a tube at 100 mc can be expected to be only one ten-thousandth of its value at 1 mc.

Lead inductance is a self-inductance of the wire connecting each tube element to its corresponding pin, cap or conneter. At high frequencies, it represents an appreciable reactance between the tube elements and the external oscillator circuit.

In the conventional grounded cathode oscillator circuits, lead inductance is of particular importance. The reason for this is illustrated in Fig. 2(A). The cathode lead inductance \( L_{cd} \) is in series with both the plate and the grid \( d \) return current. Therefore, a higher grid voltage is required to maintain the plate current, and the oscillation is less damped. The presence of cathode lead inductance \( (L_{cd}) \) causes the effective voltage between grid and cathode of the tube to have a different phase angle than that of the external applied voltage. The difference is due to the feedback voltage across \( L_{cd} \) due to plate current. The result is a conductance component in input admittance which adds to the conductance due to the transit time.

It has been shown that input conductance due to \( L_{cd} \) is

\[ g_{m} = \frac{w_{0} L_{cd} R_{pi}}{1 + \frac{w_{0} L_{cd}}{R_{pi}}} \]

where:
- \( g_{m} \) = input conductance due to \( L_{cd} \).
- \( w_{0} \) = angular velocity of oscillation (2\pif).
- \( L_{cd} \) = cathode lead inductance.
- \( R_{pi} \) = grid-to-plate resistance of the tube.

Bad effects also result from inductances of other leads as is discussed later.

To reduce the effect of lead inductance, many tubes designed for high frequency use are supplied with two or more leads and external connections from the same element. The two or more leads can then be connected together at the socket. This connects the lead inductances in parallel, thus reducing the total inductance effect to the inductance of one lead divided by the number of leads so connected. This is illustrated for a cathode lead in Fig. 2(B).

Interelectrode capacitances have a damping effect due to their relatively low resistance value at high frequencies. The charging current through these capacitances results in power loss in the resistance of the circuit and adds to the power loss in the dielectric, which is the insulating material of the tube.

Limitation by Tube of Minimum Tuned Circuit Size

The oscillating frequency is determined not by the external tuned circuit constants alone, but by the external tuned circuit plus the lead inductance and interelectrode capacitances of the tube. The combined effective tuned circuit thus reaches an irreducible minimum size (and maximum resonant frequency) when
the external tank is replaced by a direct short circuit. The effective tuned circuit is then composed of the lead inductance and the inter-electrode capacitance, as illustrated in Fig. 3. Since nothing further can be done to decrease inductance or capacitance, the tube has reached its upper frequency limit even though transit time might allow it to operate at a higher frequency.

Thus, if a tube is to oscillate at a very high frequency, its lead inductance and interelectrode capacitance must be small enough to allow resonance with some sort of external tuned circuit. Preferably, the plate, grid and cathode must be located so that a high frequency tuned circuit can be directly connected without intervening leads.

Triodes are by far the most popular tube type for high frequency oscillators, because of their low interelectrode capacitance. Types with the highest Gm are of course the most suitable. As has been previously explained, having several leads from each of the active elements is also helpful.

Influence of Circuit Construction on High Frequency Operation

Even when the vacuum tube is properly chosen, high frequency operation may be adversely affected by the character and construction of the circuit.

All kinds of circuit losses increase rapidly with frequency, i.e., if not
property controlled, may keep feed-back from being sufficient to support oscillation. Wiring must be direct and of heavy-gauge wire to combat skin effect, wherever r-f current flows. Any points of r-f voltage should be either suspended in air or mounted on low-loss material such as polystyrene or polyethylene. Although chassis grounding should be as direct as possible, it should all be done at one point in each circuit, to prevent bad effects of r-f currents in the chassis. Soldered connections must be the best possible; any tendency toward a "cold joint" or "vicia joint" can introduce extreme losses and may prevent oscillation.

Operating the Heater at Cathode E.P. Potential

The construction of modern vacuum tubes is such that there is an appreciable capacitance between the heater and the cathode (2 to 10 pf). Thus, in circuits in which the cathode is operated above ground, serious shunting of the cathode can occur through this capacitance to the grounded heater, as shown in Fig. 4. This can be overcome by isolating the heater from ground as far as r-f is concerned, and operating it in the tube at the same r-f potential as the cathode. One way of doing this is shown, at (B) in Fig. 4. An r-f choke is connected in each heater lead; the cathode r-f voltage can build up across the reactance of each choke, although the steps alternating current for the heater is allowed to pass.

Another arrangement is shown in Fig. 4(C). Here the heater is isolated from r-f ground by means of a quarter-wave resonant transmission line section in each lead. Both line sections are short-circuited to r-f at the bottom (one directly, the other through capacitor C4). This means that there is a high impedance at the other end between the inner conductor and the ground outer conductor. The heater leads are fed through these inner conductors and are thus at a high impedance to ground, thus preventing shunting of the cathode.

Types of Circuits Used

Many of the same types of circuits used at low frequencies are also popular in the VHF and UHF ranges. Such circuits as the Hartley are frequently encountered, especially in the "grounded plate" form (see Part 2 of this series). High frequency versions may be a little difficult to recognize at first, as special tank circuits and other construction are often employed.

Of particular importance is the Colpitts circuit of Fig. 5, not only because it is sometimes used itself, but mainly because it is the basis of the very popular ultrasonic circuit, which will be explained presently.

The Colpitts oscillator circuit is the same as the Hartley except that the cathode is tapped into the resonant circuit by means of a capacitor voltage divider C3-C4, instead of a tap on the coil. One advantage of this arrangement is that the two interelectrode capacitances C1 and C2 are not connected directly across parts of the coil, as they are in the Hartley. They are shunted by tuned circuit capacitors C1 and C2. The latter have large values, since they combine in series to provide the total external resonant circuit capacitance. The effect of interelectrode capacitance variation on the frequency of oscillation is thus minimized. There is one disadvantage in the Colpitts circuit when the circuit is to be tuned over a range, as in receiver local oscillators. Either C1 and C2 must be tuned together, or another capacitor must be added across all or part of the coil to provide tuning adjustment. The relative values of
C1 and C2 determine the amount of feedback, just as adjustment of the taps on the Hartley oscillator. Thus variation of either of these capacitors alone would vary feedback as well as frequency, an obviously unsatisfactory condition.

The Oscillator

The ultradirect circuit is undoubtedly the most popular of any of the circuits used for TV and UHF ranges. It is widely used as the local oscillator in commercial FM broadcast and TV broadcast receivers, because of its simplicity. The circuit is actually simply a Colpitts type in which the plate-cathode and grid-cathode interelectrode capacitances form the voltage divider across the coil. No external capacitors are then needed, although of course some form of trimmer or adjusting capacitor must usually be added across the coil, so the frequency can be set or varied.

The principle of the ultradirect is illustrated in Fig. 6, which shows how the interelectrode capacitances form the Colpitts-type voltage divider.

As with other oscillator types, any desired point is the r-f circuit can be chosen as ground, to suit convenience in the particular application. Two examples of ultradirect local oscillator circuits used in TV receivers are shown in Fig. 7. In the circuit at (A), the cathode is grounded. The plate is shorted through R1, which keeps it at r-f potential above ground. Optimum efficiency and power output would call for an r-f choke instead of R1. However, in this case, sufficient receiver injection voltage, and better stability is obtained by using the lower-priced resistor, because a voltage drop across the resistor is probably necessary anyway. In the circuit at (B), the grid is grounded to r-f through C1. This means that both the grid and the cathode must operate at above-ground r-f potential. The cathode is kept above ground by means of the cathode choke L, which allows d-c cathode current to pass through from ground to the tube.

Use of Transmission Lines as Tank Circuits

Because of the relatively high circuit losses and the effects of transmission time at high frequencies, the inherent stability of an oscillator lessens as the frequency is increased into the VHF and UHF regions. One way to compensate for this is to design the resonant (tank) circuit so it has a very high Q and thus tends to stabilize the oscillator as a whole. This can be done by using a resonant section of a transmission line as the tank circuit, instead of the ordinary coil and capacitor. For example, a quarter-wavelength section of the transmission line, shorted at one end, exhibits at the other end the characteristics of a very high Q parallel-resonant circuit. By slight adjustment of the length of the line section, it can be made to combine with circuit and tube reactances plus added tuning capacitance if desired, to resonate at the required operating frequency. An open-ended line section a half-wavelength long can be used in the same way.

An example of the use of a line section for the tank circuit of an oscillator is shown in Fig. 8. Because of the appearance of the shortened line, which is usually fashioned from a single piece bent into shape, this arrangement is often referred to as the "hairpin" oscillator. Actually the circuit is an ultradirect, and the construction is about the simplest of any practical oscillator.

This application is not limited to open wire lines, but coastal line sections also can be used. A Hartley circuit using a quarter-wave coastal section is shown in Fig. 9. The line is shortened and grounded at both ends, where the plate is also connected through C1. The circuit is thus a grounded-plate Hartley. The leads from the cathode and grid respectively are fed through the outer conductor of the line section and tapped onto the inner conductor. This simulates the connection of these leads to the tap and top respectively of a conventional coil. C3 is added for variation or adjustment of frequency. Sometimes frequency adjustment is provided by a shorting plug of metal between the inner and outer conductors, which is moved to change the electrical position of the bottom short-circuit.

Push-pull oscillator circuits have the advantage at high frequencies that the combined effective interelectrode capacitances are lower than those of each tube alone. A typical push-pull tuned-plate-tuned-grid oscillator circuit using transmission line tanks is shown in Fig. 10.
Regulated Power Supply Design

A source of well-regulated plate voltage is a prerequisite for the modern laboratory, service bench or amateur circuit. An increasing number of electronic devices, such as audio amplifiers, r.f. oscillators, etc., require a very stable voltage or current. There are many ways of achieving this, but all depend for their proper functioning upon a power supply which is free from noise and delivers a constant voltage regardless of load. Fortunately, the development of electronically regulated sources has advanced to the stage where their design and construction are well within the scope of the average experimenter. The theory, design and construction of a representative supply of this type will be outlined here, with a firm understanding of the design steps to be discussed, the reader should be able to adapt the practical supply presented here to other requirements which might exist.

Modern regulated supplies of the type to be described make available an output voltage which is continuously variable over a considerable range and which will not vary more than a fraction of one percent between normal and full-load conditions. Normal line voltage fluctuations also have little effect on output voltage. In addition, the regulation may be made of such a high order that ripple voltage is the output is almost entirely cancelled, thus eliminating the need for the so-called "brute-force" filters. This saving in weight and space helps to make the size of the supply less of a problem of the electronic regulator.

Theory of Operation

To achieve precise voltage regulation, an electronic voltage control element must be introduced in the conventional supply circuit. In modern regulated supplies, this electronically variable element takes the form of a high-current vacuum tube, usually called the "pass tube" or "regulator tube" in this application. This tube is connected in series with the load resistance across the output of the supply, as in Fig. 1. Since the resistance of the triode varies as a function of its grid voltage, this combination acts as an electronically controlled voltage divider. A small change in the regulator tube grid voltage changes the effective ratio of the divider and thus varies the voltage appearing across the output load.

The ability to vary the output voltage of the supply by a minute grid voltage change suggests that automatic voltage regulation could be accomplished by feeding any attempted output voltage fluctuation back to this grid at such a polarity as to oppose that change. In other words, if the voltage across the load in Fig. 1 attempted to rise, the grid of the pass tube (V1) should be made more negative, so that its internal resistance would increase and lower the load voltage. If the load voltage attempted to decrease, the converse action should occur.

This action is achieved by the circuit shown in simplified form in Fig. 2. Auxiliary circuitry consisting of a second vacuum tube, usually called the "control tube," and a constant voltage source such as a battery or "VR" tube is added to the circuit of Fig. 1. A sample of the output voltage is applied to the grid of the control tube by a tap on the output bleeder R1. The control tube determines the bias on the regulator tube (V1) since the load resistors (R3) for the control tube are also the bias resistors for the regulator tube. The control tube therefore performs two functions: it amplifies voltage fluctuations appearing at its grid by the output circuit, and it reverses the phase of those fluctuations so that they may be applied to the grid of the pass tube in the right direction to effect regulation. The precision of the regulator attained increases with the gain of the control tube.

\[ \text{Equation of pass tube} \]

Since, with greater gain, a small change in control tube grid voltage will cause a greater control tube current change and hence a greater change in pass tube bias. Thus, smaller attempted output voltage excursions will be corrected.

The battery of VR tube maintains the cathode of the control tube at a constant voltage above ground, and thus provides a standard reference voltage to which voltage fluctuations at the output divider (R3) are compared. The voltage at the grid of the control tube is the difference between the voltage at the output bleeder tap and the reference bias voltage provided by the VR tube. This difference voltage sets the "target" voltage to which the supply regulates. By changing the output bleeder tap with a potentiometer at R1, the regulated output voltage of the supply may be adjusted within certain limits.

Summarized briefly, the action of the electronic regulator of Fig. 2 is as follows: The position of the bleeder tap on R1 determines the output voltage level to which the supply will regulate. If the voltage across the bleeder attempts to rise above that level, the bias on the control tube (V2) becomes more negative, causing it to draw more current through its anode resistance (R6). The increased current through R6 causes the grid of the regulator tube (V1) to be driven more negative, with the result that the resistance of the regulator tube increases sufficiently to prevent the original attempted excursion of output voltage and return it to the regulated level. If the output voltage attempts to decrease, the re-
sequence of events is exactly opposite. The search is performed from top to bottom, so that ascensions are corrected for backwards.

Practical Design Considerations

With a working knowledge of the functions of all component parts, the design of regulated power supply equipment is no more complicated than that of other electronic circuits usually designed and constructed by the user.

As with any power supply design, the first step is to determine the desired output voltage and current requirements. This permits the selection of the proper power transformer, filter components, and pass tube. The supply section differs from standard design only in that considerably more voltage than the required output voltage must be provided since there is an appreciable minimum voltage drop across the regulator tube. Usually the unregulated section of the supply must furnish from 10 to 200 volts more than the desired regulated output.

For a sample design, let us suppose that we require a regulated output of about 300 volts at 25 milliamperes is required for a general utility supply. The practical circuit for such a supply is shown in Fig. 2. Knowing the current requirement, a suitable pass tube may be selected from Table I. Any triode or tetrode-connected pentode capable of passing the required current at a reasonable voltage drop may be employed. Tubes may readily be used in parallel where greater current is required or when greater plate dissipation is necessary. Such tubes as the 6A7, 6A7GT, or 6A7G tubes are usually available. For our present design, a smaller tube such as the 6A7 or 63 volt equivalent, the 6A3, will suffice.

The power transformer and filter choke must be conservatively rated for the full load current. Otherwise, the regulation of the supply will be poor. The required voltage rating for the transformer is determined by finding the sum of the voltage drops around the circuit for the condition of maximum output voltage and current. The drop across the pass tube is minimum for maximum output voltage and may be found by referring to the plate characteristic curves for the pass tube being used. For the 6A3 used in the present design, the minimum tube drop for the required load current is about 80 volts at zero bias. Actually, somewhat greater values should be assigned to provide a margin for different line voltage conditions. For the 6A3, a minimum drop of 140 volts is typical. Thus, the d.c. output of the supply section ahead of the regulator must be about 440 volts; 300 volts for the load and 140 minimum drop across the pass tube. Reference to the rectifier tube operating characteristic will indicate the d.c. voltage rating of the power transformer required to supply this voltage when a single section choke-input filter is used. With the 5U4G employed in the present design, and allowing sufficient margin for voltage drop across the choke, low line voltages, etc., a transformer delivering 550 volts each side of center-tap at 100 ma. is indicated. The choke should also be rated at 300 ma.

At this point, having selected the pass tube and determined the characteristics of the unregulated supply section, it is well to examine the pass tube operating conditions to determine if the allowed plate dissipation is being exceeded. The 6A3 is rated at 15 watts maximum dissipation. At 300 current and voltage from the supply, the drop across the pass tube estimated above was 140 volts. The plate dissipation under this condition is 140 volts, times 0.75 amperes, or 105 watt. The low voltage limit to which the supply can safely be adjusted at full current may now be determined, since the voltage drop across the pass tube, and hence its plate dissipation, is maximum at the lowest regulated output voltage. The allowable drop for 15 watts plate dissipation is now calculated as 14 volts, 0.75 amp., or 200 volts. With a total unregulated voltage of 440 volts, available, the minimum regulated output of the supply is thus 240 volts. By using a larger pass tube or several in parallel, the range of regulated voltage adjustment can be appreciably extended.

The choice of a control tube is rather arbitrary. Almost any pentode having a sharp cut-off characteristic may be used. The type most frequently employed in electronically regulated supplies is the 6SJ7, which is chosen for its low cost, ready availability, and high gain. Various types having similar characteristics may be used in applications where space is at a premium. The 6SJ7 will do nicely for the design under consideration.

Although batteries may be used for the source of control tube excitation bias voltage, the gitter grid voltage regulator tube is usually preferred. Tubes of the "VR" series give excellent life and satisfactory performance in this application. The choice of VR type, VR75, 90, 105 or 150, depends on the unregulated voltage available and the portion of this which must be reserved for drop across the load-bias resistor (R2) and the control tube. It is desirable to utilize the highest volt-

### Table I

<table>
<thead>
<tr>
<th>TUBE TYPE</th>
<th>CURRENT (MA)</th>
</tr>
</thead>
<tbody>
<tr>
<td>6A7GT</td>
<td>250</td>
</tr>
<tr>
<td>6A3</td>
<td>75</td>
</tr>
<tr>
<td>6A4G</td>
<td>75</td>
</tr>
<tr>
<td>6A6G</td>
<td>75</td>
</tr>
<tr>
<td>807*</td>
<td>80</td>
</tr>
<tr>
<td>807*</td>
<td>80</td>
</tr>
<tr>
<td>857*</td>
<td>60</td>
</tr>
<tr>
<td>857*</td>
<td>40</td>
</tr>
</tbody>
</table>

* Jumps connecting A plate through 100 ohms J push pull

PRACTICAL REGULATED SUPPLY

Fig. 3
age VR tube possible under these conditions, since this subjects the grid of the control tube to a larger portion of output voltage fluctuations. A VM/150 is sufficient for the design being discussed, since the bias developed across R2 to reduce the output voltage to minimum is only about 50 volts, as indicated by the plate curves for the 6AS. The plate load resistor (R2) is chosen to be about equal to the plate resistance of the control tube. Values between 47 and 68 megohms are typical for the 6S7.

The by-pass capacitor, C1, is usually about 25 microfarads. It provides a path for 120 cycle ripple voltages and other high frequency fluctuations between the regulated output and the grid of the control tube.

The dropping resistors R3 and R4 are designed to provide 150 volts across the VR tube at the 8 ma. minimum current required for regulation and to provide a tap for control tube screen voltage. In computing the values of these resistors, the minimum unregulated supply output voltage must be used. Allowing for 10% drop in line voltage, this would be 396 volts in the present case. The required drop is then 396 minus 150 = 246 volts. From the resistor curve, the total resistance required (R3 and R4) is 246,000 or 20,750 ohms.

The portion of this resistance between the cathode and screen of the control tube to furnish a screen voltage of 100 volts should be 100,000 or 12,500 ohms. Thus the nearest standard values of 12,000 and 18,000 ohms will suffice for R3 and R4 respectively.

The total resistance value for the output bleeder is usually about 25 megohms, made up of a 50,000 wire-wound potentiometer for the voltage output adjustment and fixed carbon resistors (R3 and R4) above and below to complete the total. The exact values of these for any particular regulated supply are most easily determined experimentally by substituting a 25 megohm potentiometer temporarily in place of R1, R5 and R6. Then, with the supply operating, the settings of the potentiometer tap for the minimum and maximum output voltages allowable under full load conditions can be determined. The potentiometer is then disconnected and the resistances measured with an ohmmeter. The resistance between the slider position for low voltage output and the ground end of the "pot" is the value for R5. Similarly, the resistance measured between the slider setting for high output voltage and the "flat" end of the potentiometer is the value for R6. The correct value for R1 is then R5 plus R6 subtracted from 25 megohms.

Ferroresonant Circuits

I f a saturable reactor is made the inductive element of a tuned circuit, either series-resonant or parallel, varying the current through the reactor will vary the resonant frequency of the combination. This is the simple basis of all ferroresonant circuits. The most efficient manifestation of this affect occurs when a large resonant frequency shift is obtained as the result of a small change in control current, usually through a separate winding. The ferroresonant effect may be utilized in various ways in electronic and electrical equipment.

Ferroresonant circuits recently have been designed to perform as amplifiers, trigger switches, flip-flops, gates and oscillators. Ferroresonant elements find application in electronic counters, computers and other digital devices, frequency dividers, signal amplifiers, and automatic control circuits. Closely related to the magnetic amplifier, the ferroresonant circuit is useless and has unlimited life. Ferroresonant elements are rugged and can be made extremely small in size. Unlike the magnetic amplifier, ferroresonant devices may be operated at high supply-voltage frequencies, enabling high-speed switching performance and high-frequency amplification.

Basic Ferroresonant Circuit

In Figure 1, a coil having a saturable core is connected in series with a conventional capacitor, a resistor, and a continuously variable a-c source. This is the basic ferroresonant circuit. The zero-current inductance of core L, and the capacitance C are chosen such that the series-resonant frequency, f, of the
The circuit can exhibit true bistability. That is, its current may have either a high or a low discrete value at a particular voltage, and this current value can be maintained indefinitely unless the operator switches it to its second value. Thus, referring to Figure 2(B), the current might have a low value X or a high value Y along the inscribed non-linear load line. The current will not hold along the unstable, negative slope, AB, but can be made to shift from some point along OA to a point along BC, quickly traversing AB and locking along BC. If the ferroresonant element is operated in a suitable circuit conduction may be triggered back and forth between the two stable states. This action suits the ferroresonant element ideally as a binary device.

**Ferroresonant Flip-Flops**

One of the most promising and already exploited applications of small ferroresonant elements is as flip-flops in digital computers, counters, and frequency dividers.

**A second, coupled winding might be added to the simple circuit of Figure 1.** This modification is shown in Figure 3. Here, the control winding actually consists of two coils (L1 and L2) which are connected series-back-to-back to isolate the TRIGGER INPUT terminals from the ac supply. A dc trigger pulse of either polarity applied to the control winding then would reduce momentarily the inductance of L1 and allow the current to jump from X (Figure 2B) to Y. The current then would latch at the second value and would be unaffected by further pulses applied to the control winding, since the high current at point X would be sufficient to maintain the core saturated.

The fact that the current is this simple circuit cannot be returned to
as initial level by alternate pulses, but only by a momentary interruption of the a-c supply line prevents use of the circuit directly as a flip-flop even though it is bistable.

Figure 3 is an arrangement of two ferroresonant elements in parallel. Each has a separate control coil. This arrangement provides good flip-flop action. The circuit is due to

loosen. Here, L_C1 and L_C2 correspond to the simple elements shown previously in Figure 1. Capacitor C1 provides a series reactance which is common to both ferroresonant legs. The control winding for the left leg consists of coils L1 and L2 connected in series-backing to prevent transmission of the a-c supply energy back to the INPUT 1 terminals. Similarly, the control winding of the right leg consists of L3 and L4 connected in series-backing to prevent a-c coupling back to the INPUT 2 terminals.

With the proper magnitude of inductance at C3, only one ferroresonant leg can conduct in the high-current state at one time. If both legs should attempt to operate at this level simultaneously at high current, the in

Figure 4
creased voltage drop at point X would reduce the leg drops (R<sub>L</sub> and E<sub>L</sub>C<sub>L</sub>) below the resonant value and neither could latch at high current. When both legs drop to low-current conduction (or attempt to do so), the drop across C<sub>L</sub> falls and the voltage at point X rises to such a high value that either the left or right leg will “fire.”

In order to examine operation of this circuit, assume that the left leg (L<sub>L2</sub>C<sub>L2</sub>) is conducting high and that the right leg (L<sub>L1</sub>C<sub>L1</sub>) is conducting low. Output 1 will be high (TRUE or 1 in binary notation) and output 2 will be low (binary FALSE or 0). A pulse applied to INPUT 1 will have no effect, since the core already is saturated in this leg of the circuit. Coil L<sub>C</sub>, however, is carrying low current and its core is not saturated. A pulse applied to INPUT 2 therefore will lower the inductance of L<sub>C</sub>, saturating the core and moving the L<sub>C</sub>C<sub>L</sub> leg into resonance. The high current then will reduce the voltage at point X momentarily to such an extent that L<sub>C</sub>C<sub>L</sub>, detuned from resonance and drops to the low current condition. High conduction then shifts to the right leg, and OUTPUT 2 becomes high. Thus, the two outputs always are of opposite phase and change states with respect to each other.

For single-input operation, INPUT 1 and INPUT 2 may be connected together and the triggering pulse thus applied simultaneously to L<sub>L1</sub>C<sub>L1</sub> and L<sub>L2</sub>C<sub>L2</sub>. Successive pulses then will cause high conduction to flip back and forth between the two legs, and a given pair of OUTPUT terminals will have high output during only half of the number of input pulses.

The circuit thus acts as a frequency halver. Appreciable power gain is obtained. Rithsahan has pointed out that by providing control coils of many turns, the required triggering current can be made very small. One flip-flop thus can drive others without intermediate driver stages. In this way, it is comparatively simple to set up labelless binary counter rings counters, and similar circuits. Rithsahan has had a ring of 82 stages operating satisfactorily and states that about 50 stages would appear to be the upper limit.

Ande from being tubeless, simple, and capable of extreme miniaturization, ferroresonant flip-flops with small cores of thin material may be operated at high power-supply frequencies (1 Mc and higher) allowing rapid switching. Lown mentions that the entire switching operation takes place in a period equal to approximately 5 cycles of the supply frequency. Another considerable advantage of the ferroresonant flip-flop is its low power requirement. This results from the use of reactive components in the circuit. Furthermore, although high current is shifting from one leg to the other, the load on the power supply is constant. Another advantage is the low heat radiation which results from low power dissipation and the absence of tube filaments.
Triax of International Business Machines Corp. has patented an improved ferroresonant flip-flop, shown in basic arrangement in Figure 5. This circuit employs a pair of saturating windings (L₁ and L₂) on each core and cross-couples them. A low lagging current flows in the less saturated leg and a high, leading current in the more saturated leg. The patent claims that the cross coupling increases the differences between the two current magnitudes, resulting in a wide gap. The combined resistance, R, in Figure 5 serves the same purpose as the common capacitor, C₁, in Figure 4.

The main saturable inductor windings are L₁ and L₂. Coils L₃ and L₄ are the auxiliary windings. When the left leg is saturated, high leading current in L₄ produces flux in the core of the right leg to induce an e. m. f. in L₃. This voltage opposes the lagging current in L₃. The net result of this action is a reduction of the current in the right leg and an increase in the current in the left leg. This is in the way that the gap between the values of voltage drop across capacitors C₁ and C₂ is widened. Outputs are taken from across these two capacitors.

Additional Applications

Mackay has developed several interesting circuits of a highly practical nature in which ferroresonant elements are combined with the non-linearity of tungsten lamp filaments to obtain relaxation oscillations and the intermittent operation of lamps. This scheme has been described also in the popular literature.

Figures 6 and 7 show two of Mackay's circuits. In Figure 6, L₁ is the primary winding of a small filament transformer (Sunray P6355) and serves well enough as a saturable reactor at the current levels involved. C₁ is a 10-microfarad capacitor (non-electrolytic) and R₁ is a 100-watt lamp. The latter exhibits high resistance (high current) when its filament is cold, and high resistance (low current) when hot. The L₁ and C₁ combination behaves like the simple ferroresonant circuit shown in Figure 1 when the current is high enough to start saturation of the coil L₁. The circuit is operated from the 115-volt ac power line through a variable autotransformer such as a Variac.

When the circuit is switched on, the filament is cold and its resistance low. The lamp accordingly lights from the correspondingly high current flow and the ferroresonant circuit operates at its high-current point. As the filament heats, its resistance increases, the current accordingly drops and will snap (flip-flop fashion) to the low-current point on the ferroresonant response curve. The lamp therefore extinguishes. As the filament then cools, its resistance decreases, current again rises and soon snaps back to the high-level point to repeat the cycle of operations. When the input voltage is adjusted to a certain critical point, the lamp flashes on and off at a regular rate as the result of this ferroresonant action. Mackay gives a 5-second ON rate with the circuit constants shown here.

The circuit shown in Figure 7 will be seen to resemble the ferroresonant flip-flop of Figure 4, minus control (trigger) windings. It consists of two "blinder" circuits of the type just described, with the addition of the common capacitor, C₁. In this circuit, L₁ and L₂ are the primary windings of small filament transformers, as before, and C₂ is a 10-microfarad non-electrolytic capacitor, C₃ and C₄ each are 12 microfarads, also non-electrolytic. Each lamp is of 100-watt size. Both lamps cannot light simultaneously because the large voltage drop across C₁, due to the resulting high current flow, would reduce the voltage across each leg of the circuit to a value much lower than the ferroresonant width and insufficient to light the lamps. However, one lamp always will have enough voltage for the lamp in that leg to glow. When that lamp extinguishes, the one in the other leg ignites. Mackay has operated a number of lamps through the common capacitor and observed that the lamps flash cyclically.

An obvious application of this low-frequency type of ferroresonant circuit would be as a high-intensity blinder in path and tunnel lights as traffic signal beacons, railroad signals, etc., where at moving parts, relay contacts, and the like would be eliminated and continuous, unattended operation assured.
SECTION III
COMPONENTS

Fixed Capacitors in Modern Circuitry

No other electrical component is called upon to perform such a wide variety of functions in electronic circuits as the capacitor. Most of these applications are based upon the ability of the condenser to differentiate between electrical currents of various frequencies. Such applications include: clamping, ripple filtering, r-f and s-o noise by-passing, coupling, frequency determination, R-C timing, and energy storage. Because of the varied requirements of these uses, fixed capacitors are made in many types and sizes, each especially engineered to fulfill a specific application or function. An important part of modern circuit design is therefore the choice of the proper capacitor for the circuit application at hand. In many cases, the success or failure of the design will actually depend upon this choice. The radio engineer, experimenter, and amateur must therefore have a firm background in capacitor design and application.

Probably the most direct route to a mastery of the "safe and sane" use of capacitors is to establish a thorough understanding of the characteristics and limitations of each general type. The choice of the proper type for each circuit application then becomes merely a matter of following good engineering practice. For this reason, we will commence with a discussion of the basic types of fixed capacitors which are encountered in electronic circuitry.

![Capacitor Equivalent Circuit](image)

Since a capacitor is fundamentally two metallic conducting sheets isolated by a suitable dielectric material, the basic types are classified according to the type of dielectric used. They include:

- Air-Dielectric Capacitors
- Micro Capacitors
- Ceramic Capacitors
- Paper Capacitors
- Electrolytic Capacitors

Just as all inductances have distributed capacity and resistance, and everyday resistors have some inductance and "end-to-end" capacitance, practical condensers are not perfect capacitances. All have a certain amount of residual capacitance associated with the leads and plates, and also a finite value of resistance called the "insertion resistance". Thus, the equivalent circuit of any capacitor can be considered as in Fig. 1. The magnitudes of these unwanted characteristics vary though wide limits as a function of mechanical design and type of insulation or "improvement" used, and must be considered along with such other characteristics as capacitance value, voltage and current ratings, temperature coefficient, stability, etc., in selecting a condenser for a particular job. The actual choice is usually a compromise between mechanical and electrical perfection on one hand, and the dictates of economy, space, and the practical requirements of the application on the other.

The Air-Dielectric Capacitor

From the standpoint of low losses (high capacitance) and constancy of capacity value, the most nearly ideal capacitors are built with air (or vacuum) as the dielectric between the plates. Such capacitors are not perfect, however, for although air is a perfect dielectric having zero power factor, some losses arise due to dielectric hysteresis in the insulating material used to support the plates. Charging currents flowing in the leads and plates cause additional power losses and give rise to some residual reactance.

The air-dielectric condenser occupies much more volume for a given capacitance and is usually more expensive than any of the other general types. The reasons for this are apparent from an inspection of one of the simpler empirical formulas for the capacitance between parallel plates whose dimensions are large compared with the spacing between them, so that "fringing" may be neglected:

$$C = \varepsilon \frac{ad}{d^2}$$

where:
- $C$ is the dielectric constant of the material between plates.
- $\varepsilon$ is the area of the plates in $\text{cm}^2$.
- $d$ is the distance between the plates in $\text{cm}$.

From this it is seen that the capacitance is directly proportional to the dielectric constant and the plate area, and inversely proportional to the spacing. Since the dielectric constant of air is only 1.0, but is greater than unity for all other insulating materials used in condenser construction, it is obvious that air can be used in capacitors to achieve a given capacitance. In addition, the dielectric strength of air is considerably lower than that of the other dielectrics, so that greater electrode spacings are necessary for a given working voltage. As a result, the volume occupied by an air-dielectric condenser will be at least 500 times greater than that of a comparable capacitor using a high grade mica dielectric.
Because of these factors, air as a dielectric is used only to a very limited extent in fixed capacitors, such as in certain laboratory capacitance standards. Fixed capacitors using vacuum or an inert gas under pressure are used to a greater extent, since the breakdown voltage is increased about four to ten times thereby. Air dielectric variable capacitors are, of course, widely used for tuning r.f. circuits because of their mechanical simplicity.

Mica Capacitors.

Mica is widely used as the insulating material in capacitors manufactured primarily for r.f. applications. The mica capacitor is characterized by low power factor, high puncture voltage, good stability, high insulation resistance, and reasonable cost. As mentioned above, the size for a given capacity is considerably smaller than that of a comparable air-di- electric capacitor. Due to the stacked construction usually employed, the insulation is quite low. A common construction is illustrated in Fig. 2. The plates consist of metal foil sandwiched between thin sheets of mica dielectric material. The sides of alternate foil strips extend beyond the mica sheets at opposite ends of the stack and each group is clamped together and connected to a lead. Thus, the central currents which flow in each plate do so through a relatively short, broad path. Therefore, the inductance is low, being mostly that contributed by the wire leads.

Typical Mica Capacitor Construction

Mica capacitors are used in a multitude of electronic applications where a high degree of capacitance excellence is required. Such uses include: r.f. fixed tuned circuits, c.r. by-passing, c.r. coupling, d.c. blocking, r.f. neutralizing, r.f. filtering a.c. tone control, a.c. degenerative feedback, a.c. coupling where high insulation resistance is important (as in section c.c.-coupled amplifiers), and many others.

In radio frequency applications, mica capacitors are rated according to r.f. current handling capability as well as maximum instantaneous voltage. The observation of both of these ratings are equally important in practical use.

Exciseive r.f. current results in capacitance heating, which, in turn, causes intrinsic dielectric losses, capacitance deviation, and lowered breakdown voltage. The effect is thus cumulative. The r.f. current through a capacitor in any given application can be determined by connecting a suitable r.f. thermocouple in series with it.

In applications where stability of capacitance value is important, as in tuned circuits, c.r. filters, and other critical circuits, capacitance of the "filled mica" variety are used. These units have extensive capacitance stability and low temperature coefficients. These excellent characteristics are obtained by depositing a silver coating on the opposite surfaces of mica foils and "airing" this assembly at high temperature to form highly conducting metal "plating" in intimate contact with the mica. The variable factor of stacking pressure is thus drastically reduced with correspondingly improved stability.

Use of Temperature Compensating Capacitor

High quality mica units are manufactured with either positive, zero, or negative temperature coefficients of capacitance. Capacity of this type can be used for temperature compensation in tuned LC circuits in which low frequency drift with ambient temperature changes becomes intolerable. By such means, a.c.-excited c.r. oscillators having frequency stability comparable to crystal controlled oscillators can be built. Stabilized oscillators of this type are used for various local oscillators, among them, power oscillators where voltage control is impractical, etc. An example of the application of temperature compensating mica capacitors is given in Fig. 3. Here it is desired to maintain the LC product (and hence the frequency of an r.f. oscillator "tank" circuit) at a constant value over a wide temperature range. This may be accomplished by determining the approximate temperature coefficient of the uncompensated circuit in terms of capacitance deviation in parts per million per degree Centigrade. This coefficient will usually be positive with common circuit elements, i.e., the frequency increases with increasing temperature. Temperature compensation then consists of the selection of a capacitor having a negative temperature coefficient approximately equal to the positive characteristic of the other circuit elements. Thus with all circuit elements subjected to the same ambient temperature change, frequency "drift" is compensated. A trick frequency correction to by circuit designers consists of placing the compensating capacitor at a location in the equivalent where a temperature gradient exists, such as near a vacuum tube. A "vernier control" of temperature compensation is thus obtained by adjusting the position of the capacitor within this gradient by trial and error until a point of best frequency stability is located.

The Ceramic Capacitor

Another type of condenser which is in some cases comparable to the mica capacitor in electrical characteristics uses a ceramic as the diele- tric material. A typical design is shown in Fig. 4. The capacitor plates are deposited on the inner and outer surfaces of a ceramic tube with connecting leads at either end. This unit is then sealed in a second ceramic tube and the whole assembly is wax impregnated for moisture proofing.

Ceramic capacitors are manufactured in a wide variety of characteristics and are used in the same applications as mica and glass. The type of the ceramic have very high dielectric constants, the volume efficiency (micromicrofarads, cubic inch) is high. Titanium dioxide ceramics, for instance, are used extensively and are available in high dielectric constants (600- 1000), low losses and low temperature coefficients. Since the temperature coefficient can be controlled by the ceramic mixture, tolerances ranging from essentially zero to high negative values of temperature coefficients are available for temperature compensation. Due to the varied type of construction, tubular ceramic capacitors have low values of residual inductance.

One grade of ceramic capacitor is used interchangeably with mica capacitors in critical r.f. circuits, while having a lower quality variety which has
very high volume efficiencies but poor stability, is used for general purpose applications such as bypassing. Ceramic tubular capacitors are usually more expensive than equivalent mica units. However, disk type ceramic capacitors are just as expensive as equivalent mica capacitors and are sold on a "guaranteed minimum value" basis. Disk ceramic is used in high frequency by-pass applications only.

Paper Capacitors

Capacitors using wax or oil impregnated paper dielectric are employed extensively in d.c., audio and low frequency r.f. applications where high capacitance per unit volume and low cost is required. They are characterized by generally poorer electrical characteristics than mica or ceramic capacitors, including: higher power factor, lower temperature coefficients, lower operating voltages, higher inductance and shorter life. These factors depend to a large extent upon the type of impregnant used, the purity of the impregnant, the method of construction, and the casing employed.

Wax is used as the impregnant in a large variety of utility capacitors for the lower voltage ratings, where small size and economy are important. The tubular capacitors used in receiver audio, by-pass, and by-pass work are examples. Moisture absorption shortens the life of unsealed wax capacitors to some extent, as does high ambient temperature. Careful drying and oil immersion in a desiccator will extend the life of wax capacitors, if used properly, for a reasonable length of time.

Typical paper condensers have temperature coefficients of capacitance approximately ten times larger than high grade mica capacitors, such as the silvered-mica types. Power factors are greater by at least one order of magnitude and capacitances are larger, especially in the types using foil or rolled foil construction. In such units the correct taps are at the ends of the rolled foil plates. In paper capacitors of advanced design, residual inductance is minimized by the use of the extended electrode construction, in which electrical contact is made at the edges of the rolled electrodes, so that charged current paths are short.

In applications where a wide range of frequencies must be effectively bypassed, as in the TV line filter shown in Fig. 5, a high capacitance paper capacitor may be used in parallel with a small mica unit. Otherwise, the residual inductance of the paper condenser may make it ineffective as a by-pass for the high r.f. frequencies. Another by-passing device used in video l.f. amplifier design consists of using capacitors which are self resonant at a frequency to be by-passed. A value of capacitance is chosen, which is series resonant with the inherent inductance of the capacitor and its leads. This type of single-frequency by-passing is very effective.

The Electrolytic Condenser

The familiar electrolytic capacitor is the "work horse" of the receiver power supply filter fields. These units have extremely high volume efficiencies, occupying only about 15% of the space required for equivalent paper capacitors. The cost per microfarad is also very low. For these reasons, although inferior in most other respects to the other types, the electrolytic capacitor is extensively used for filter and by-pass applications.

An electrolytic capacitor may be made either by immersing two aluminum or copper plates in a solution such as ammonium boric at sodium phosphate ("dry" electrolytic). A forming voltage applied between the plates deposits a film of aluminum oxide on the positive plate. See Fig. 6. This film is the dielectric material of the capacitor. Because it is extremely thin—being only .000025 inch thick in some cases—the capacitance per unit area is very high. For the same reason, the operating voltage of the unit is limited to about 400 volts. Electrolytics, however, may be used in series for higher voltages with the use of the usual voltage equalizing resistors shunting each unit, as point used with mica and paper capacitors which have higher insulating resistances.

Proper Use of By-Pass Condensers

In modern electronic circuits the 0.1 microfarad capacitor is used more frequently for the function known as "by-passing" than for any other single application. The selection of a capacitance of the proper type and value for a given job is an important part of circuit design. Such critical performance characteristics as frequency response, phase distortion, circuit stability, and freedom from parasitic oscillations are determined by the by-passing used. This discussion is intended to provide a review of this subject for the benefit of the amateur, experimenter, young enginner, or anyone who has been puzzled by the problem of what by-pass to use for a specific purpose. The factors underlying the choice of capacitors in practical circuits will be pointed out by the use of examples.

"By-passing" can be defined as providing a short, low impedance path
around certain circuit components for electrical currents of some frequencies, while maintaining a high impedance path for other frequencies. The circuit designer is repeatedly confronted with the need of components having this property of passing currents of a desired periodicity while excluding others. Actually, both inductances and capacitors qualify under this definition because of the frequency discriminating action of these simple filters. An inductance, or "choker," may be considered to be a low frequency bypass element since it presents a high impedance path for dc and low frequencies while presenting a high reactive impedance for high frequencies. The condenser, on the hand, is a simple high-pass filter, having a high reactance at low frequencies and becoming more nearly a short circuit as frequency is increased. It is when this latter property is used to provide a "detector" around some part of a circuit that the term by-passing is most commonly employed.

\[
10^6 \times 10^7 \times 10^8 \times 10^9 \times 10^{10}
\]

\[
\text{Frequency (c/s)}
\]

\[
X_L (\text{ohms})
\]

\[
10^1 \times 10^2 \times 10^3 \times 10^4 \times 10^5
\]

\[
X_C (\text{ohms})
\]

\[
10^4 \times 10^5 \times 10^6 \times 10^7 \times 10^8
\]

\[
\text{Frequency (c/s)}
\]

\[
\text{FIG. 1}
\]

For a capacitor to function as an effective by-pass, its impedance must be much lower than the impedance of the circuit element being by-passed. Of course, the reactive impedance of a capacitor of any value is easily calculated from the given frequency from the basic expression:

\[
X_c = \frac{1}{2\pi f C}
\]

Where:
- \(X_c\) is the capacitive reactance in ohms
- \(f\) is the frequency in cycles per second
- \(C\) is the capacitance in farads

Provided that \(X_c \gg X_L\), \(X_c\) is low reactance.

Needless to say, this relationship is of constant use in designing proper by-pass circuits. It shows that the reactance of a given unit decreases with frequency or that, for a given frequency, a value of capacitance can be chosen to give any desired value of capacitive reactance. To aid in visualizing this function, we have plotted the reactance of a 0.001 mfd. condenser versus frequency in Fig. 1.

**Cathode Resistor By-Passing**

The most frequent use of the by-pass condenser is illustrated in Fig. 2, where the capacitor is used as a cathode resistor by-pass. The necessity for this is obvious when the characteristics of the circuit are considered. As is well known, any vacuum tube stage which uses cathode bias exhibits strong regeneration if the signal current is allowed to flow through the bias resistor. This is so, as the a.c. component of the plate current flowing through the bias resistor develops a voltage drop across it during signal peaks which increases the bias applied to the grid of the tube. This has the effect of reducing the signal voltage on the tube grid and thus reducing the stage gain and introducing phase distortion and other undesirable results.

In Fig. 2, this regenerative effect is prevented by shunting the cathode bias resistor with a capacitor which by-passes the a.c. signal component around it. Let us now consider the requirements placed upon this capacitor.

Assume that the stage depicted in Fig. 2 is an audio amplifier intended to work over the frequency range of 200 to 5000 c.p.s. and that the cathode resistor recommended for the tube type used is 200 ohms. A by-pass capacitor must be provided across this resistance which will prevent most of the audio frequency plate signal current from flowing through it. Since the reactive impedance of the condenser becomes lower with increasing frequency, as shown by Eq. 1, one which is satisfactory at the low frequency end of the desired range will do for the entire range. Therefore, in the present example, a capacitor which effectively by-passes the 100 ohms cathode resistor at 200 c.p.s. should be adequate. Most circuit designers consider a ratio of bias resistance to by-pass reactance of about 10-to-1 to be a safe rule-of-thumb for most work. With this ratio more than 99% of the total a.c. current flows through the by-pass condenser. Ratios up to 30-to-1 may be used in high fidelity amplifier work where space and economical considerations permit, however.

Assuming a by-pass ratio of 10-to-1 to be sufficient, a capacitor having a reactance of one-tenth the resistance of the bias resistor at 200 c.p.s. is necessary. By rewriting Eq. 1 to solve for a value of capacitance having a reactance of 30 ohms, an answer of 26 mfd. or 0.026 is obtained. Therefore, the nearest standard value of 25 microfarads would be used. An electrolytic condenser is usually used in this application since leakage resistance is not important in this case and these units are cheap and economical. The capacitor must be rated for a working voltage greater than the maximum bias voltage developed. This may be obtained from Ohm's Law, using the bias resistance and the maximum d.c. current which flows through it. For pentodes, this means both the plate and screen current, and for classes of amplification other than Class A requires the maximum signal current. A voltage rating of 25 or 50 volts usually sufficient for capacitor by-passing.

**Drive C C FOR Audio**

\[
\text{FIG. 3}
\]

In the example discussed above, the 100 ohms cathode resistor could have been made large with little limit, without detrimental effects on circuit performance. Circuits exist, however, in which change in an upward direction to the capacitance which can be used to by-pass the resistance in the cathode circuit. As an example, consider the cathode-modulated Class C i.f. amplifier shown in Fig. 3. Here the condenser is required to by-pass r.f. around the modulated transformer. Otherwise regeneration may result from feedback into the grid bias circuit. However, if the cathode by-pass is made too large, the modulation frequencies will be shunted to ground. A value of capacitance must be chosen which has very low reactance at the
carrier frequency, but a high one at the highest modulation frequency. Fortunately, this is easily done in this case because of the wide difficulty in the frequencies involved; a 0.02 µf microfarad condenser has a reactance of about 2.0 ohms at a 1 frequency of 10 Mc, but almost 16,000 ohms at 5000 c.p.s. A good paper or ceramic condenser of low inductance would be used in this application.

Of course, not all cathode bias re-actors must be by-passed. In many high fidelity, audio amplifiers, and television L.F. amplifiers controlled amounts of negative feedback are introduced to improve the overall performance. In such cases, the loss of gain is compensated by adding extra stages. Cathode by-passing is also omitted in Class A push-pull amplifiers, since the a.c. signal components of both tones flowing in the resistor are out of phase and cancel out.

Screen By-Passing The screen element of triode and pentode electron tubes must be effect-ively by-passed to ground for all signal voltages present. This is neces-sary to prevent degeneration of a type very similar to that discussed above. For example, consider the screen voltage of an L F. stage shown in Fig. 4. Here the screen voltage is derived from the plate supply through a dropping resistor. If the screen component is allowed to pass through the stage, the output of the stage will be reduced. For this reason, a by-pass condenser is used to ground the screen for signal voltages without interfering with the application of the L.F. screen voltage. If the screen by-passing is imperfect at any frequency, the response of the amplifier will fall off there or it will oscillate. It is common practice to make the screen by-pass reactance small compared with the cathode-to-screen impedance. This is obtained by dividing the screen voltage by the screen current.

Mica or ceramic condensers are used in voltage rating from 100 or 1000 µf microfarads for radio frequencies, while high quality paper units and electrolytics are used for audio screen by-passing. As in cathode re-actor by-passing, certain circuits require screen by-passing suffi-ciently heavy to ground the screen for L.F. but not for audio frequencies. A typical example of such a stage would be the plate and screen modulated Class C amplifier shown in Fig. 5. In this circuit the screen voltage must vary with the modulation and so should not be by-passed for audio frequencies. A 0.02 µf microfarad condenser is sufficient in most cases and does not result in a loss of "highs".

Plate Circuit By-Passing As in the cathode and screen circu-its discussed above, any impedance in the plate circuit of a vacuum tube stage is common to another stage, or another part of the same stage, can cause feedback and instability if not properly by-passed. The reasons for this are obvious from Fig. 6.

Plate voltage for two stages of an amplifier may be taken from the same power supply and no decoupling is employed. The internal impend-ance of the power supply is represent- ed by Rs. Since diode plate current is allowed to flow through Rs, a voltage drop is developed across Rs which is subtracted from the plate circuit of the preceding stage via the plate load. The signal voltage is then fed to the grid circuit of the second stage, with the result that oscil-lation will occur if the stage gain is high enough.

Instability due to plate circuit feed-back is prevented by the use of se-rially decoupling filters consisting of series isolating resistors and by-pass con-densers, as shown in Fig. 6b. Such filters are important in order to prevent high frequency feedback between stages. A small signal voltage (Es) is therefore developed across the power supply impedance and travels down the plate lead to the preceding stage. The function of the decoupling filter Rs and C2 is to greatly attenuate this signal since they divide it in the ratio of their impedances. Thus if the reactive impedance of C2 is only 1 ohm and the resistance of Rs is 1000 ohms, the feedback signal is divided by that ratio so that only 1/1000th of the voltage developed across the power supply impedance is applied to the preceding stage. Of course, the L.F. plate voltage is unaffected except for a small IR drop across Rs. In cases where this drop couldn't be tolerated, an inductor could be used in place of Rs. Several such RC or LC de-coupling filters are sometimes used in series with feedback to partic-uilarly troublesome.

By-Passing Precautions By-Passing a vacuum tube circuit, including high gain amplifiers and video amplifiers, must be done with extreme care to avoid common imped-ance which introduces feedback. The
Non-Linear Resistors

A FEW electronic devices are distinguished by their non-linear current-voltage characteristics of a magnitude sufficiently great to affect performance. Often this phenomenon is a characteristic. For example, non-linearity is observed in the plate characteristic of a vacuum tube under certain operating conditions, the extreme of a d.h.d. device, intermodulation products operated in the region of saturation, and in plate capacitors having special dielectric properties (diode amplifiers). There are only a few examples. The non-linearity of certain 2-terminal devices such as tubes also may be employed in modifying the operating characteristics of some circuits. These latter non-linear resistors do not obey the simple relationship \( R = \frac{V}{I} \) of Ohm's Law.

Simple non-linear resistors are used in oscillators, wave shaping networks, voltage regulators, current regulators, constant-output transformers, and voltage dividers, voltage-sensitive circuits, transistors, frequency multipliers, surge suppressors, etc. Their use in electronics is increasing as new requirements for non-linear current-voltage response arise.

The general characteristics of some common 2-terminal non-linear resistors of several classes will be described here. In some instances, as will be seen, these devices have other prime uses and their applications as non-linear resistors are secondary. Typical applications will be shown.

Thyratrons are applied in the electric power field where a number of years before entering electronics. Thyrite resistors having leads of \#20 wire spooled with a filter, then pressed and fired at high temperatures. They are fabricated in the form of pigmented rods (identical to "radio resistors"). They, radios, and others. Small units suitable for electronic applications are supplied up to 10 watts power ratings (continuous).

ILLUSTRATING USE OF BY-PASS GROUND POINTS

FIG. 7

grip-flop meter or other absorption indicating device. The easiest length of the wire lead to be used in the circuit must be used in this measurement for precise results. This is illustrated by the fact that the resonant frequency of a tubular 31mil wire of about 111 is about 15m. However, if the leads are minimized as much...
The non-linearity of the Thyrite resistor is expressed by $I = kV^n$, where $I$ is the instantaneous alternative or direct current (amperes), $V$ the instantaneous applied voltage, $k$ a constant (amperes at 1 volt), and $n$ an exponent between 3.5 and 7 governed by the manufacturing process. Figure 1 shows a set of typical Thyrite current-voltage curves for several types of G.E. Thyrite resistors. The curves for a conventional 1 megohm linear resistor is plotted for comparison. From these curves, it may be seen that large current changes result from small applied-voltage changes.

Figure 2 is a plot of resistance vs. current for resistor B from Figure 1. Note in Figure 2 that a change of 20,000 to 1 in current flowing through this resistor changes the resistance of the latter approximately 37%. Thyrite resistors have the advantage of operating in both a.c. and d.c. circuits. Any rectification effects are negligible. (Figure 2 shows a typical static positive-negative conduction curve). High-frequency a.c. operation of Thyrite is possible, the limiting factor appearing to be capacitance. It should be noted, however, that in a.c. operation the non-linearity of the Thyrite volt-ampere characteristic causes distortion of the current waveform. Figure 4 shows the distorted current wave accompanying a sine wave of applied voltage. Observe that odd-ordered harmonics are prevalent. This phenomenon is utilized in simple frequency multipliers and harmonic accentuators.

The temperature coefficient of Thyrite resistance is negative in sign and varies from $-0.4$ to $-0.73$ percent per degree Centigrade in the range 0 to 100°C.

Figure 5 shows several typical circuits utilizing the properties of Thyrite resistors. Figure 5(A) is a simple voltage regulator or preamplifier for smoothing variations in supply voltage. The fluctuating input voltage produces alternating current which flows through the Thyrite resistor (T) and a linear limiting resistor (R) in
FIGURE 6

**STATIC DC CHARACTERISTIC OF THERMISTOR**

Fluctuations in the resulting voltage drop across the Thermistor resistor are considerably lower in amplitude than those in the supply voltage due to the non-linear E/I relationship in the Thermistor. It should be noted that a voltage reduction is unavoidable because of the Thermistor's active role in the circuit. In the expander circuit (Figure 5B), the expander action is secured. Small fluctuations in applied voltage produce large fluctuations in current through the Thermistor resistor. These fluctuations in turn produce a fluctuating output-voltage drop across R. Since the resistance of R must be small with respect to that of T, voltage divider action between the Thermistor and linear resistor legs of the circuit produces a drop in output level. Several TR sections may be cascaded, as shown in Figure 5(C), to secure additional expander action, at the expense of course of further voltage division in each section.

Figure 5(D) shows a potentiometer or voltage divider with Thermistor sections. The non-linear E/I characteristic of the Thermistor section yields a nearly constant output voltage at each tap, although supply and load currents are variable.

For efficient operation of the circuits shown in Figures 5(A) and (D), the Thermistor current must be high with respect to the output load current. In Figures 5(B) and (C), current in resistor R must be high with respect to output load current.

Figure 5(E) shows how a Thermistor resistor can be connected into a circuit to accentuate harmonics and act as a simple frequency multiplier. Alternatively, current is fed into the circuit and undergoes distortion in passing through the Thermistor. The tuned circuit, LC, is adjusted to the desired multiple frequency. It has been shown already in Figure 4 that odd harmonics are favored by this type of operation. The Thermistor frequency multiplier thus is most practical for tripling, quintupling, etc. The Thermistor resistor is a dissipative element, however, and its insertion into a circuit in the manner shown in Figure 5(E) results in some power loss. In applications where a considerable amount of power is available, the relative simplicity of the Thermistor frequency multiplier can offset its unavoidable power absorption.

**FIGURE 7**

**TIME-DELAY ACTION OF THERMISTOR**

Circuit, LC, is adjusted to the desired multiple frequency. It has been shown already in Figure 4 that odd harmonics are favored by this type of operation. The Thermistor frequency multiplier thus is most practical for tripling, quintupling, etc. The Thermistor resistor is a dissipative element, however, and its insertion into a circuit in the manner shown in Figure 5(E) results in some power loss. In applications where a considerable amount of power is available, the relative simplicity of the Thermistor frequency multiplier can offset its unavoidable power absorption.

**FIGURE 8**

**TYPICAL THERMISTOR CIRCUITS**

**FIGURE 9**

**LOCK-OUT SWITCHING CIRCUIT**

46
Thermistors

The thermistor, a product of manufacture by Bell Telephone Laboratories and manufactured by Western Electric Company, is another interesting non-linear resistance device. Its action results from internal heating effects in special materials.

Thermistors basically are thermal-sensitive resistance devices. They are manufactured in the shape of rods, discs, heads, washers, and balls and are made of various semiconductors.

Like Thyratrons, the thermistor can be used with either a.c. or d.c.

Figures 6 and 7 show two important response curves describing thermistor action. From Figure 6, it is seen that the voltage drop across the thermistor increases non-linearly and rapidly, with inverse flow up to a point beyond which the rate of increase falls. Finally, a peak is reached and beyond this latter point, the voltage drop decreases with increasing current, displaying negative resistance. An interesting side observation is that this negative-resistance property has been utilized to obtain low-frequency undamped oscillation and amplification with thermistors.

In Figure 7, the plot shows how at a particular applied voltage, internal heating causes the magnitude of the thermistor current to vary as a function of time. This property has been utilized in various simple time-delay devices.

Figure 8 shows several simple circuits employing thermistors. In all examples, Figure 8(A), a small current-limiting resistor, R, is indicated. Figure 8(A) is a time-delay circuit, i.e., relay based upon the action of the choke by the curve in Figure 7. Some seconds after the switch, S, is closed, the circuit current rises to a value high enough to close the relay. The relay interval depends upon thermistor characteristics and supply voltage level, and can be adjusted to some extent by means of linear series resistance.

Figure 9(B) is the circuit at a regulator for supply-voltage variations and is somewhat similar to the Thyristor voltage regulator. Its operation is based upon the non-linearity of the thermistor which results in smaller variations in thermistor voltage drop than the fluctuations occurring in supply voltage and current.

Function of the limiter circuit, shown in Figure 8(C), is similar to that of the voltage regulator, amplitude excursions in the input signal being reduced in the output without clipping or slewing action.

Operation of the thermistor for the circuit shown in Figure 8(D), is the inverse of that of the limiter. The thermistor and load resistor are interchanged in position, output being taken across the resistor. A smaller signal voltage change produces a linear thermistor current change and a large voltage change across the output resistor.

A voltage divider takes place in the circuits shown in Figures 8(B), 8(C), and 8(D), as the result of the potential divider action between the thermistor and the linear series resistor. Since the action of this effect is the same as that of a voltage divider, the potential divider action is reduced in the output.

Figure 8(E) shows a best-watch-type switching circuit employing thermistors. In each of the circuits, R is a load resistor or represents some device, such as a relay, which is to be actuated by current flowing through the associated thermistor. The supply voltage and the value of the linear series resistor R are chosen such that the resistor will support the current of only one led before its voltage drop becomes excessive.

When one switch (say, S1) is closed, the associated thermistor "breaks down" allowing current to flow through and operate the associated device, D1. This lowers the voltage at the inside of R, so that no other thermistor can "fire." Only after S1 is opened, can either of the other circuit leads be operated. Operation of any one thermistor leg thus locks out all of the other legs. An arrangement of devices having similar voltage characteristics could be connected across a single voltage pair, but with only one device operative at a time.

Fluorescent Devices

The tungsten-filament irradiated lamp is fairly well known as a non-linear resistor in which current change lags a corresponding change in applied voltage. Up to the point at which heating effects begin to evidence themselves, the filament volt-ampere characteristic is linear, or very nearly so. The non-linear region of lamp-filament resistance has been utilized in voltage-stabilization bridges, simple regulators, and allied devices.

In a common application, the lamp-type resistor is used as an automatic regulator of degeneration voltage in low-distortion, R/C-tuned oscillators. A typical circuit is shown in Figure 9.

In Figure 9, the lamp (R1) is the cathode resistor of the first pentode. The output voltage, V, is coupled from the output of a frequency-selective R/C network to the grid circuit of V1. A portion of this current also flows through the thermistor R1 and the lamp, R2, establishes a negative feedback voltage across the latter. The lamp resistance changes as feedback changes, the feedback voltage across the latter. The lamp resistance changes as feedback changes, the feedback voltage across the latter.

Tungsten Lamp (R2) as Stabilizing Resistor in Audio Oscillator

Figure 9
inverse feedback voltage across R2, and this degeneration in turn reduces the amplitude. The opposite also is true; at weak oscillation amplitudes, there are lesser amounts of degeneration, and gain through the two-tube circuit automatically rises. The net result is uniform amplitude of oscillation.

Thermistors also are used occasionally in some RC-tuned oscillators to stabilize oscillation amplitude.

Small filamentary, low-current fuses exhibit a type of non-linearity somewhat similar to that of the tungsten filament. Figure 10 shows the static d.c. voltmeter characteristic of a sample Type BAG 10-milliampere line fuse. In this instance, the response is linear from zero up to the 0.8 V, 4 ma. point. Beyond this, the non-linearity is apparent.

When d.c. biased to a point within the square-law region of their non-linearity, such fuses often are used as bolometer-type detectors in microwave work. This provides an extremely simple and inexpensive detector at frequencies up to many hundreds of megacycles.

Diode-Type Resistor

Non-linearity in the forward conduction characteristic of the germanium diode invalidates this simple component to use as a non-linear resistor in applications within its current capabilities. While the reverse conduction (back-current) characteristics of the diode also is non-linear, it does not in general offer the same possibilities of application that are available with the forward conduction?

Figure 11 shows a plot of forward resistance vs applied voltage for a high-conductance-type germanium diode. Here, the polarity of the applied voltage is such that the anode of the diode is positive. Diodes may be connected in series, parallel, series-parallel, and parallel-series to obtain many attractive non-linear resistance effects.

Various portions of the forward voltmeter characteristic of the germanium diode exhibit square-law, logarithmic, and finally approximately linear relationships between E and I. By operating the diode as a desired one of these regions, the particular corresponding portions of the curve can be utilized to correct or modify the R/I characteristic of another circuit. For example, a linear milliammeter can be converted into a square-law instrument by using the forward resistance of the diode in the meter circuit (multiplier).

Diodes suffer somewhat in comparison with other terminal non-linear resistors because the diode is a rectifier. This limits application in some cases to direct-current use only. However, small a.c. signals may be superimposed upon a d.c. forward bias current applied to the diode, the two currents being so proportioned that the net diode voltage never becomes zero or negative. Diodes also are relatively low-current devices, as compared with some other non-linear resistors.

Like the Thyrite resistor, the forward-conducting diode is capable of distorting a.c. current waveform and occasionally is used to accentuate harmonics. The requirement is that the diode current magnitude be small enough to operate the diode in its most non-linear region. Thus, the single series connection of a diode in the plate or grid lead of an oscillator or amplifier can accentuate harmonic content of the current waveform when this type of operation is required. An example is the distortion of waveform of a standard-frequency oscillator to produce high-order harmonics for calibration purposes.

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48
The extensive development in electronics during World War II saw the return to active service of many circuits and devices which had previously been considered obsolete. Such things as the superregenerative detector, the magnetron oscillator, and the rotary spark gap were "recovered," subjected to further development, and ultimately found important applications in military radio and radar equipment.

Not the least important among these "rediscovered" devices is the crystal detector. For, although the silicon, germanium, or iron pyrites crystal, with its ever-present "catwhisker," was a household item during the early days of broadcast radio, it was ultimately replaced by the vacuum tube. Then, when transit time and noise limitations ruled out the vacuum tube as a detector or mixer for microwave radar, the crystal detector was again rejected.

As a result of this war-time rejuvenation, the clumsy crystal detector of old has emerged as the remarkably dependable, efficient, and compact unit known as the "semiconductor diode." These units are finding extensive use in many types of circuitry. The amateur radio operator and experimenter has been especially quick to adapt the versatile, economical crystal to an ever-widening variety of uses. With that in mind, we present some of the more recent uses for crystal diodes in amateur projects and enumerate the constructional features which make such devices a good addition to any amateur's kit.

**Constructional Features**

The present popularity of the crystal diode is due to fact that its miniaturization makes it a practical circuit element which is capable of outperforming the vacuum tube in many instances. The elimination of the vexing process of searching for a suitable "catwhisker" by the use of a fixed rectifying contact makes it possible to package the crystal in a very compact capsule. The unit is also quite rugged and moisture-proof, since the capsule containing the diode elements is impregnated with a wax filler.

The general construction of a point-contact rectifier is illustrated in Fig. 1. The essential elements are the small "nucleus" of a specially processed semiconductor material, a fine tungsten-wire catwhisker that is sharpened to a point at the end in contact with the semiconductor, and an insulating body or capsule which holds these two parts in rigid contact and provides external electrical connections to them. There are several constructional variations which meet these specifications, each intended for a specific type of application. Some crystal diodes for r.f. applications are designed for insertion into a "socket" consisting of spring contact fingers, while others intended primarily for low-frequency and video work, are equipped with pigtail leads and have the general appearance of small resistors or tubular capacitors. More recently, units which are hermetically sealed in glass envelopes have become available.

**Electrical Advantages**

In addition to the mechanical advantages of small size and ruggedness enumerated above, the point-contact rectifier has several important electrical attributes which make it preferable to vacuum tube rectifiers for many uses. Of course, one obvious advantage is the elimination of filament power consumption. This adds to the overall efficiency and makes it easy to use the crystal rectifier where the tube must have low capacity to ground.

At extremely high radio frequencies, the two electrical characteristics which are most responsible for the usefulness of the crystal diode are the low inter-electrode capacitance and the short transit time. Since the sharpened point of the tungsten wire makes contact with the semiconductor over a very small area, the capacitance of modern crystal rectifiers may be less than one microfarad. Transit time is negligible in most crystals because the rectifying "barrier layer" through which electrons must flow between the semiconductor and the metal contact may be only one millimicron of a centimeter in thickness—much closer than it is possible to space the electrodes of a vacuum tube. For these reasons, the crystal has been used to detect r.f. energy well into the millimeter wave region.

**Crystal Types**

Crystal diodes may be divided into two major categories: the high sensitivity type for high frequency detector and mixer applications, and the high-back-voltage variety which serves as the overall efficiency rectifier and second detector. The high sensitivity type includes such silicon crystals, while the high-back-voltage types use a germanium semiconductor.

**TABLE I**

<table>
<thead>
<tr>
<th>No.</th>
<th>Use</th>
<th>Upper Freq. (MHz)</th>
<th>Remarks</th>
</tr>
</thead>
<tbody>
<tr>
<td>1101</td>
<td>Mixer (6)</td>
<td>10,000</td>
<td>Improved IN-4A</td>
</tr>
<tr>
<td>1102</td>
<td>(bar)</td>
<td></td>
<td>High Sensitive</td>
</tr>
<tr>
<td>1103</td>
<td>(bar)</td>
<td></td>
<td>High Sensitive</td>
</tr>
<tr>
<td>1104</td>
<td>(bar)</td>
<td></td>
<td>High Sensitive</td>
</tr>
<tr>
<td>1105</td>
<td>(bar)</td>
<td></td>
<td>High Sensitive</td>
</tr>
<tr>
<td>1106</td>
<td>(bar)</td>
<td></td>
<td>High Sensitive</td>
</tr>
<tr>
<td>1107</td>
<td>(bar)</td>
<td></td>
<td>High Sensitive</td>
</tr>
<tr>
<td>1108</td>
<td>(bar)</td>
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<td>High Sensitive</td>
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<tr>
<td>1109</td>
<td>(bar)</td>
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<td>High Sensitive</td>
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<tr>
<td>1110</td>
<td>(bar)</td>
<td></td>
<td>High Sensitive</td>
</tr>
<tr>
<td>1111</td>
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<td></td>
<td>High Sensitive</td>
</tr>
<tr>
<td>1112</td>
<td>(bar)</td>
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<td>1113</td>
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<td>1114</td>
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<td></td>
<td>High Sensitive</td>
</tr>
<tr>
<td>1115</td>
<td>(bar)</td>
<td></td>
<td>High Sensitive</td>
</tr>
<tr>
<td>1116</td>
<td>(bar)</td>
<td></td>
<td>High Sensitive</td>
</tr>
</tbody>
</table>

(6) Denotes silicon crystal
(10) Denotes germanium crystal

49
Table 1 lists some of the various types which are, or have been commercially available, and gives the recommended upper and lower frequency limit. As a general rule, crystal rectifiers can be used with good rectification efficiency and performance at any frequency below this limit.

To date, the high-back-voltage family of crystal diodes has enjoyed the greatest popularity in amateur radio applications. This is due to the fact that a greater variety of low-frequency use for crystals as vacuum tube substitutes have been found. The 1N34 germanium diode, in particular, is used in dozens of circuit applications where a vacuum-tube diode such as the 6H6 or the 5AL5 would otherwise be employed. The high-back-voltage types are usually limited to frequencies below 100 mc. because the rectification efficiency of germanium falls off above this frequency.

With the growing use of the microwave amateur bands, it is expected that the silicon diode will be used to a much greater extent in amateur gear. At frequencies above the 420 mc. band the crystal mixer has decided advantages over vacuum tubes from noise considerations. Crystal mixers have been produced which have noise figures approaching the theoretical optimum—the noise which would be produced in an equivalent resistor. Of course, since the crystal contributes no signal gain, the first i-f stage succeeding it must also have a good noise figure. The low-noise cascade amplifier is ideal for this purpose.

Applications of Germanium Crystals

Most of the uses of germanium diodes is amateur work to date have been as rectifier devices in r-f detecting devices such as the TVI harmonic checker for locating spurious transmitter radiations which was described in Section 11 under Television Interference Filters. Detecting instruments of this kind, when tightly coupled and used with a microammeter, will indicate r-f energy of only a few microwatts.

Other indicating devices which have basic circuits very similar to the harmonic checker are the crystal field-strength meter, the absorption wave-meter, and the modulation monitor. An interesting variation of this circuit is the carrier failure alarm shown in Fig. 2. It may be employed to indicate the presence of r-f power at the antenna of "phone" transmitters and so prevent "lost" transmissions due to the failure of the antenna change-over relay, transmission line, or other components. The alarm relay used must be of a sensitive type which will pull up at about one milli-ampere. The alarm indicator may be a signal light or a buzzer.

Another family of circuits which commonly use germanium diodes are the directional coupler type of standing-wave indicators and power monitors. Fig. 3 is the schematic of a typical resistance bridge standing-wave indicator which illustrates the principle of these very useful instruments. This simple circuit can be used to measure the standing wave ratio existing on a transmission line and thus determine the impedance of the antenna or other terminating device at its end. It consists of a resistance bridge which is balanced when the cable terminating resistance (R1) is equal to R2. Under this condition, no current flows through the germanium crystal and the meter reading will be zero. If, however, the line resistance does not equal R1, the bridge is unbalanced and a current which is proportional to the degree of impedance mismatch flows through the crystal and is indicated on the meter. The value of R1 must be equal to the surge impedance of the transmission line with which the bridge is to be used. Calibration may be accomplished by connecting various non-inductive resistors of known, value greater than R1 across the output transmission line and noting the meter deflection. The voltage standing-wave ratio is then equal to R2/R1.

A graph of VSWR versus meter reading may be plotted, or a special scale on the meter may be marked directly in VSWR. Before readings are taken, the meter deflection is adjusted to full scale by open-circuiting the...
by varying the filter resistance. Other mixer circuits, such as full-wave and bridge connection, may be used.

Still another type of application for the germanium diode family makes use of the negative resistance portion of the blocking voltage curve. This unique characteristic, which occurs when a negative voltage on the cathode is blocked by a current on the anode, has enabled the IN34 and other high-ohm voltage diodes to function as sine-wave oscillators and voltage regulators. Fig. 5 depicts basic circuit for these uses.

Use of Silicon Crystals

For frequencies above 100 mc, a silicon crystal should be used in r.f. rectifying devices such as the hermetic crystal detector, field-strength meter, carrier failure alarm, and standing-wave indicator discussed above. The IN222 is an instrument rectifier which is well suited to such applications.

The silicon crystal diode is also employed in high frequency circuits where high sensitivity is required. The circuit of a typical h.f. superhetodyne crystal mixer is shown in Fig. 6. The type 42 condenser (C1) must have a low value of capacitance but a high value of the r.f. frequency. Tuning is done by varying the capacitance between the inner and outer conductors of the coaxial circuit.

Proper Electronic Wiring Techniques

THE old adage that warns that "a chain is no stronger than its weakest link" is nowhere truer than in the art of electronic wiring. Especially in today's advanced circuitry, where marginalization of reliability is the keynote, proper wiring is a prerequisite. Where the function of a guided missile, the proper functioning of an airport blind landing system, or the final answer of an electronic computer may depend upon any one of thousands of tiny soldered connections, meticulous attention must be paid to such small details. Even in the research and development field, it is impossible to estimate how many important experiments have failed because trivial circuit trouble masked the desired results. And in less glamorous applications, such as the telephone industry and the radio and television manufacturing and servicing fields, the extra effort expended in producing dependable wiring has been found to pay dividends. For these reasons, a working knowledge of the proper methods of producing a neat, dependable wiring job are required of every technician and engineer in the field of electronics.

The basic steps in wiring an electronic device are essentially the same regardless of whether the unit is an entire telephone switchboard or a tiny 4" x 6" control panel. They are as follows:

(a) Mounting the circuit component.
(b) Wiring and cabling.
(c) Connecting and lead dress.
(d) Visual inspection.
(e) Electrical inspection.

We will discuss here these wiring operations in some detail, with special emphasis on (b) and (c).

Mounting Circuit Components

Here it is assumed that the circuit layout has been arranged so that the components are located in the unit's best positions which give short lead lengths for critical circuits, minimize the effects of hum fields radiated by chokes and transformers, and places temperature sensitive components in the most desirable environmental locations. This layout is usually arrived at by using a "mock-up" of the unit and parts and moving the parts around until a suitable arrangement is found. The layout engineer should also provide for the use of the components which have the most suitable terminal arrangements for that particular job.
equipment, and quality audio equipment, all wiring and component parts are kept parallel to the sides of the chassis, so that the finished job presents a neat, "right-angled" appearance. Where groups of wires take the same path, they are "cabled" or "laced" together with wired cord. The comparison between a chassis wired in this manner and one in which point-to-point wiring is employed, is shown in Fig. 2.

The proper use of the techniques of lacing wire forms is a sure mark of the skilled wiring technician. The lacing stitch illustrated in Fig. 3 is standard throughout the Bell System and is simple and efficient. Note that the parts of the lacing cord which run parallel to the Wires emerge from under the part that encircles the wire. This stitch is self-locking and will remain tight even when the stitching on either side of it is cut. Wires which run from one chassis to another in a relay rack, as well as within the chassis, are protected by a much neater appearance and are stronger and more dependable if laid together in this manner to form a compact cable. Lacing also provides an index to the proper location of wires on a terminal strip or other circuit component by "breaking-out" each wire at its proper location with a separate strip, as illustrated in Fig. 4. This allows the wires to be disconnected at any time and subsequently returned to the proper terminals.

If more than one unit of a kind is to be wired with cable wires, or if an identical neat job is one in which the production technique of using a "forming boom" is employed. This consists of a large wooden board (Fig. 5) on which pegs or nails are laid to represent the shape the laced wire form must take. Holes may also be drilled through the board at the location of each terminal to which the wires must be connected. The location of holes and terminals on the forming board are determined by carefully measuring the corresponding distances between the components on the chassis or between chassis. Then the individual wires are run between the proper points on the board as indicated by the wiring diagram. Un liable labor can perform complete wiring in this manner with few mistakes, since the forming board can be clearly marked with numbers or colored coding to indicate the proper points of attachment for each wire as well as its proper routing and sequence in the wiring operation. The end of each wire is anchored in the board by wrapping it.
around a nail or peg placed in a position representing the location of the terminal it will ultimately be connected to. Enough excess should be allowed to permit gripping the insulation to the proper length. If holes are used in the board to "tie" the wires through at the desired location, the points of attachment will be on the back side of the forming board.

When all of the wires have been run on the forming board, they are cabled together while still on the board. Here they are usually more accessible than in the chassis, making for a faster and better job. The wires may also be stripped of their insulation and readied for connecting after the cabling is complete so that they are held firmly in the position they will assume in the chassis.

Connecting and Lead Dress

Connecting is the operation of electrically attaching the wires to the component terminals. The type of connection employed depends on the kind of terminals provided on the circuit components. There may be binding screws, soldering lugs, or directly soldered connections. Soldering to some type is employed in all of these methods but in none of them should the strength of the joint depend upon the strength of the solder. A firm mechanical attachment must be made between the wire and the terminal which is independent of the solder.

In the binding screw type of connection where soldering lugs are not used, the wire should be "tinmed" with solder and then formed into a clockwise loop around the screw, so that it will tighten as the screw is tightened. If soldering lugs are used, the grippers on the lug are cramped around the insulation on the wire for mechanical strength and the bare tip of the wire is soldered to the lug. For directly soldered connections, the well-cleaned wire is wound once around the pre-tinned terminal for strength before soldering. Winding the wire around the terminal more than once usually makes subsequent removal of the connection difficult.

The two secrets of producing good soldered connections are the preparation of the joint to be soldered and the maintenance of the soldering iron. Of course, the materials used play an important part also. Typical wire of the "push-back" variety should be used. To be possible for easy of "tinning" and connecting, the solder for all radio wiring should have a very fine flow in the vise and be a high tin content allow.

If enamelled wire is used, the insulation must be stripped back and the enamel scraped off to expose the bare metal. Otherwise, it is desirable to solder joint and then to paint. The method used is to strip the insulation from back-up wire desired upon the kind used. Most types are conveniently stripped by crushing the insulation with long nose pliers and dressing the frayed end with the diagonal. For the tougher types of insulation, such as the cellulose acetate treated kinds, a stripping tool is required. Care must be exercised to prevent "nickering" the wire during stripping and cleaning, as this frequently results in a broken connection later. As mentioned above, when the wire is done by running the wires between holes in a forming board, the wires can all be skimmed at one time by reversing the board and removing the insulation to the right length.

To produce a clean, dependable soldered connection, the soldering iron may be of a type well adapted to the job and carefully maintained. A versatile type of soldering iron tip for general connections is shown in Fig. 6. The tip should be dressed frequently with a fine and tinned wire, still bright only on the surface isolated in the drawing. The remaining parts of the tip should be allowed to oxidize. An iron tinned in this manner can be used in close places where it is necessary for it to touch other connections without flowing them since the oxide acts as a heat insulator.

The tinned portion of the soldering iron tip should be wiped on a cotton or leather pad attached to the soldering iron stand before each use. This removes excess solder and "slug" and delays motion of the tip. The operation greatly improves the quality of the job.

In the actual mechanics of soldering a connection, the professional operator does not apply the solder to the iron and transfer it to the joint being soldered. Little flux reaches a joint soldered in this manner. Instead, the well cleansed iron is applied to the junction of the wire and terminal until both have been heated to near the soldering temperature. Then the solder is applied between the wire and the connection until it flows freely around the junction. A rotary motion of the wet which "rockets" the iron tip on the joint will serve to work the solder into the joint. The finished connection should be round.
ed and smooth and the solder should have a bright, shiny appearance. If the joint is disturbed before the molten solder has solidified, the solder will look dull and "soggy" and must be melted again.

After the wires have been connected, the leads are dressed to improve the appearance of the wiring job. The insulation on each wire is pushed up against the terminal so that no exposed wire is visible. A small amount of slack is left in each wire to facilitate future reconnecting and to remove tension from the soldered joint.

**Visual and Electrical Inspection**

Before a newly wired circuit is placed in actual operation, it should be subjected to both a visual and electrical inspection so that any wiring errors can be eliminated before the application of operating voltages causes damage to circuit components.

The first inspection should consist of a thorough visual examination of the wiring to detect shorts caused by blobs of solder or loose wire ends between terminals, poorly soldered joints, broken wires or terminals, etc. When such defects have been rectified, the circuit should be given an electrical continuity test with an ohmmeter or lamp or buzzer tester. Short circuits between wires and to ground should be tested for as well as electrical continuity between points indicated on the circuit diagram.
SECTION IV
WAVE FORMS AND WAVE SHAPING

Non-Sinusoidal Wave Forms
Part 1, Passive Wave-Shaping Circuits

The recent rapid advance of such developments as radar, a-electronic television, pulse modulation systems, electronic navigational aids, computers, and other electronic devices has focused attention to an ever-increasing extent upon the “core and feeding” of non-sinusoidal electrical impulses of special shapes. As used in modern terminology, a non-sinusoidal voltage or current may be described as one whose variation with time does not satisfy the equation:

\[ v(t) = v_m \sin \omega t \]

This admittedly “back-handed” way of defining what a non-sinusoidal impulse is not perhaps simpler and more concise than a lengthy definition of what is. Fig. 1 depicts graphically several of the more common types of voltage waveforms which may be encountered in modern timing circuits. A more or less general discussion of the generation, shaping, amplification, and use of some of these waveforms, using techniques available to the circuit engineer, will comprise the bulk of this discussion.

Part 2 of this discussion is confined to the passive circuits which may be used to form non-sinusoidal waves, while the succeeding part will treat self-sustaining generators and other wave forming circuits which employ non-linear elements such as vacuum tubes.

Perhaps the simplest types of waveform shapers, or “pulse shapers” as they are called, are the circuits composed of various combinations of the passive, linear network elements, namely, resistors, capacitors, and inductors. Consider first for example the circuit of Fig. 2, in which are shown a hypothetical signal generator capable of producing any of the waveforms of Fig. 1, a pulse shaper in which AA and BB denote terminals to which may be connected any of the passive circuit elements so that various network configurations may be studied, and an oscilloscope for viewing the output voltage of the pulse shaper. The series combination of resistance and capacitance shown connected to the terminals in this case is usually termed an "RC differentiator" or "pulse sharpener," since the output voltage measured across the

(a) SINE WAVE
(b) SQUARE WAVE
(c) SAWTOOTH WAVE
(d) RECTIFIED SINE WAVE
(e) TRAPEZOIDAL WAVE
(f) EXPONENTIAL WAVE
(g) PULSE
resistor is, within certain limits, closely proportional to the time derivative of the input voltage. Other
basic configurations of passive elements which are of importance are shown in Fig. 3.

We will first consider the R-C differentiator in some detail. It may be seen from a mathematical con-
tideration that if the applied voltage in Fig. 2 is a pure sine wave, such as may be obtained from a good audio oscillator, the steady state output of the shaper will also be a pure sin-
usoid of identical frequency and waveform. In general, the only difference between the two will be their
relative amplitudes and phases; the output leading the input by a phase angle given by:

$$\beta = \tan^{-1} \left( \frac{-1}{ta} \right)$$

and reduced in amplitude as shown by the equation:

$$V_{out} = V_{in} \cdot e^{-ta} \sqrt{1 + \frac{1}{1 + ta^2}}$$

Inspection of Eqs. 2 and 3 shows that the phase angle and attenuation both decrease with increasing frequency.
It should also be noted that the phase angle may approach but never quite reach 90 electrical degrees for a single R-C circuit when a phase shift of 90 degrees requ-
since a phase shift of 90 degrees requ-
since a phase shift of 90 degrees requ-
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A set of somewhat similar equa-
tions governing the phase and attenu-
ation characteristics of the so-called R-L differentiators (Fig. 3a) for which the output voltage is propor-
tional to the time derivative of the input current, either than the input voltage as it is the case in the previous example, may be found in the literature. The phase shifting characteristic of resis-
tor-capacitor and resistor-inductor net-
works is made use of where an ac-
curately predetermined time or phase difference is required between trig-
ger pulses. A typical phase shifter circuit which may be used to accom-
plish this is shown in Fig. 4. With this device, relative phase differences of
approximately 180 electrical degrees between input and either of the outputs or
nearly 360 degrees between outputs, may be readily achieved. Thus, al-
though the phase shift for a single R-C or R-L network is limited to somewhat less than 90 degrees, it is
possible to increase the total shift to any desired value by cascading two or
more networks.

The "integrating" circuits (Fig. 3b and 3c) are so arranged because of the fact that the output voltage is propor-
tional to the time integral of either the circuit current or the applied
voltage (3c). It should be mentioned in passing that the phase and attenuation characteristics of integ-
ator circuits, which are essentially low-pass filtering circuits, are essen-
tial, like those of the differentiators (or high-pass filters) mentioned above. The major difference between
the two is that integrators display (1) a leading, rather than a trailing
phase angle, and (2) attenuation and phase angle which increase rather than de-
crease with increasing frequency.

Another type of differentiator cir-
cuit which, although little used in the
past, is of sufficient interest to war-
nant a brief discussion here is the trans-
formative or "mutual inductance" type shown in Fig. 5. This circuit offers several distinct advantages over
the previously described types. As in the standard R-L differentiator, the output voltage is again propor-
tional to the time derivative of the input current as shown by the equa-
tion:

$$V_{out} = \frac{d}{dt} \int_{0}^{t} V_{in} dt$$

where M (the mutual induc-
tance) is the proportionality factor relating the
coefficient of coupling and the pri-
mary-secondary inductions. The
transformer, like the R-C differenti-
ator, is an a-c coupled device, and as such provides more flexibility than does the R-L circuit, in which the
input and output are capacitively coupled.
Thus, it is possible to use the transformers to differentiator at a coup-
ling device between two circuits oper-
atating at different d-c levels (as in the plate and grid circuits of amplifier stages) without resorting to compi-
licated biasing arrangements. Another advantage to the circuit is the com-
pactness of case with which polarity re-
versal and voltage step-up may be
achieved, if desired.

An important concept which will aid in gaining a clearer understanding of the behavior of the passive wave-
shaping circuits of Fig. 3 is that of the time constant, T. Simply defined, T is the time in seconds required for an
uncharged capacitor C to charge through a resistor R to 63.2% of the
applied voltage V. Conversely, for a charged condenser discharging through a resistance, T is equal to the
time required for the voltage to decay exponentially to 37% of its initial value. In a similar manner, T may be
defined for an R-L circuit as the time required for the current to fall to 63.2% of its maximum value E/R, or
as the time in which the current will fall to 37% of the initial value.

The equations:

$$t = \frac{R}{V}$$

are the mathematical conventions which have been adopted.
Let us now consider the response of the circuits of Fig. 3 to non-sinusoidal waveforms. It will be remembered that for these circuits the input and output waveforms were identical under conditions of zero wave excitation, as mentioned above. Such is not the case for the non-sinusoidal waveforms, however. If, for example, a square wave of voltage (Fig. 1a) is applied to the input of either type of differentiator shown in Fig. 3, the nature of the output voltage developed depends on the value of the time constant $T$ in relation to the period $T$ occupied by one cycle of the input voltage. If the ratio $T/t$ is small, the output under these conditions will appear as a succession of alternatively positive and negative pulses which are narrow near the peaks but broader at the base as in Fig. 6a.

The output of an integrator circuit, on the other hand, with similar square wave excitation, and time constant, will resemble Fig. 6b. As may be inferred by comparison of Figs. 6a and 6b with the original square wave, it can be said in a more or less qualitative manner that the differentiator circuits transmit only the higher order frequencies contained in a complex waveform, while the integrator networks, conversely, pass only the lower or frequency components.

The qualitative analysis of the preceding paragraph may be extended to include waveforms other than the square wave. For example, consider the sawtooth waveform shown in Fig. 1c. The output of a differentiator with sawtooth excitation will, for large values of $T$, resemble the input in shape. As $T$ decreases, the waveform will in general be distended as shown by Fig. 6c. This distortion of a given complex waveform by passive networks has been recognized by Waldichuk, Rockett, and others as providing a rapid method of checking circuit and amplifier characteristics. Since the sawtooth waveform contains both even and odd harmonics of the fundamental, as compared with the square wave which contains only odd harmonics, the use of the former in such applications will result in a much more complete picture of amplifier performance.

Integrator and differentiator circuits have received their widest applications in the broadcast television receiver field. Here they serve the function of separating the high frequency horizontal pulses and the low frequency vertical pulses from the composite "sync" signal which contains both horizontal and vertical synchronization information. The time constants of the sync separators must be so adjusted that none of the horizontal sync pulses appear at the output of the integrator, and none of the vertical sync pulses appear at the output of the differentiator. In this application, the integrators are usually made up of two or three cascaded sections in order to ensure more perfect separation and also to provide comparative freedom from non-linear electrical distortions such as auto-ignition interference.

Part 2 will discuss typical generators for production of non-sinusoidal waveforms as well as wave-shaping networks employing non-linear elements.

**Non-Sinusoidal Wave Forms**

Part 2, Generators and Non-Linear Shapers
"drive": generators such as the squaring amplifiers, limiters and asciters. The self-excited generators require no input other than the usual d.c. operating potential (plus a synchronizing or triggering signal if desired for interposed stability), and are self-sustaining as long as the d.c. power is applied. They operate on the rather simple principle that any system—electrical, acoustical, or mechanical—having two stable states of equilibrium may oscillate between these states if a sufficient amount of its output energy is fed back to the input in the correct phase. Thus, it will be readily seen that the actuation of a watch movement satisfies these conditions and is therefore a mechanical relaxation oscillator. The primary source of energy in this case is the potential energy stored in the spring.

Among the electronic pulse generators to be discussed is the familiar gas tube relaxation oscillator first used in the "dark ages" of electronics as an oscilloscope time-base generator. Briefly, the sequence of action in such a device, shown diagrammatically in Fig. 1, is as follows. After switch closure, condenser C charges exponentially through resistor R, the condenser voltage at any time being given by:

\[ V(t) = E \left(1 - e^{-t/T}\right) \]

Where
- \( V(t) \) is the instantaneous condenser voltage
- \( E \) is the battery voltage
- \( t \) is the time constant, equal to the product of the time constant and the resistance
- \( t \) is the time in seconds, after switch closure
- \( t \) is the response in amperes
- \( C \) is the capacitance in farads

If the gas tube were not in the circuit, the condenser voltage would in time be equal to the full battery voltage. However, when a gas tube having an "ignition" potential somewhat lower than the battery voltage is connected as shown in Fig. 1, the condenser voltage at some point on the charging curve will be sufficient to initiate the gas, rendering it conductive. In this condition the tube serves as a discharge path for the charge stored on the condenser. The condenser voltage will decay exponentially toward zero until it reaches the extinction voltage of the gas tube permitting the circuit to take place, whereupon the condenser voltage again charges toward the ignition potential and the cycle repeats. The output voltage for large values of the RC time constant is the distorted sawtooth shown as the solid line of Fig. 2. Also shown (dotted lines) is the output voltage for identical battery voltage and gas tube conditions but with a small time constant (RC). With this simple relaxation generator, the output must be limited to about 5 or 10 percent of the supply voltage if a linear sawtooth is desired.

A pulse of very short duration may be obtained from the gas tube type of generator by inserting a resistor into the anode circuit of the pyrotron. The voltage pulse across this resistor is, of course, proportional to the discharge current pulse of the condenser, and may be made of short duration by the proper choice of tube and circuit parameters.

The blocking oscillator, Fig. 3, has been widely used in television receivers as a pulse generator for intricately timing the sweep generator. Here again the modus operandi is the relaxation principle, with an iron core transformer serving as the feedback coupling mechanism. The transformer must of necessity be so phased that an increase in plate current causes an increase in grid voltage in the positive direction, and vice versa. Under these conditions the slightest fluctuation of plate current is sufficient to drive the tube either into cutoff or plate-saturation, depending on the polarity of the initial disturbance.
For a positive-going disturbance, the plate current will increase until the rate of rise of plate current decreases sufficiently to make the rate of change of net grid voltage negative, at which time the plate current falls to zero. During the fall of plate current, as well as during the rise, the transformer phasing dictates that the rate of change of current shall be extremely rapid. Since the grid drains appreciable current during a portion of the cycle, once the plate current has fallen to zero, it will remain cut off until the accumulated negative charge on the grid condenser (and consequently the net grid voltage) becomes sufficiently small to again permit the flow of plate current. As may be seen, the cycle is repetitive and self-sustaining.

Fig. 4 depicts typical grid and plate voltage waveforms observed in blocking oscillators. The grid voltage may be used as the trigger pulse to fire a discharge-tube type of television sweep generator which will be discussed later, or the plate output may be used directly as the shaped sweep waveform. Both methods have been used satisfactorily in commercial television receivers.

Synchronization of a blocking oscillator may be accomplished by injecting a positive sync pulse into the grid circuit as shown in Fig. 3. For best results, the free-running frequency of the blocking oscillator (controlled by R₁ in the above figure) should be slightly lower than the frequency of the sync pulses, although synchronization with multiples or submultiples is possible.

Several modifications of the free-running blocking oscillator should also be mentioned. For example, if sufficient negative bias is applied to the grid, the oscillator may be prevented. This circuit, called a "driven blocking oscillator," will oscillate for only one cycle each time a trigger pulse is applied. Another useful variation is the "positive grid blocking oscillator," in which the grid is operated with a slight positive bias to reduce random pulse-to-pulse time "jitter".

Closely related to the ordinary blocking oscillator is the generator first referred to in British literature as the "squeezing" oscillator. In the circuit, the feedback mechanism is an R₁ transformer. The action of this type is similar to that of the blocking oscillator, with the significant difference that a burst of plate current normally occurs in the feedback oscillator, whereas there may be several cycles of oscillations generated before sufficient charge accumulates on the grid condenser to cut the tube off. This action, incidentally, is identical with that occurring in superregenerative detectors. Squeezing oscillators have been tried as rather primitive radar transmitters, but the synchronization problems involved have precluded their wide usage.

Another circuit which may be used as a non-sinusoidal wave generator is the multivibrator. In its simplest form (Fig. 5), it may be thought of as a two-stage, resistance-coupled amplifier whose output and input are regenerative coupled. The circuit operation may be readily understood by recognizing that two stable states are possible, tube No. 1 conducting when tube No. 2 is cut off, and vice versa. The tube that is cut off will remain in that condition until the negative voltage charge accumulated on its grid condenser during the pulse conduction period has decayed sufficiently to permit plate current to flow. The light coupling makes the build-up of current extremely rapid. When this occurs, the phasing is so arranged that the other tube will be cut off, and the cycle repeats. Like the blocking oscillator, the multivibrator may be synchronized by injecting a sync pulse into either grid circuit.

As mentioned above for the blocking oscillator, the free-running multivibrator may be altered in several ways for varying applications. If, for example, the circuit is arranged so that one tube is permanently biased to cutoff, and a positive trigger pulse is applied to the grid, the circuit will complete only one cycle of multivibrator action and return to the most stable state—that of the permanently biased cutoff condition. This type of multivibrator, called a Kipp relay, may be further modified by removing the condensers and providing conductive feedback paths. With this circuit, the tube that is cut off will remain in that condition until triggered by a positive pulse. The circuit will flip to the other equilibrium condition with the other tube cut off, and will remain that way until it is triggered similarly. The shift occurs only once for each triggering impulse applied.

The output of a symmetrical multivibrator will be the square wave illustrated in Fig. 10 of Part I. The term "symmetrical" implies that the values of resistance and capacitance in the grid and plate circuits of one of the tubes are equal, part for part, to those of the other tube. For this circuit, the frequency of oscillation is given by:

\[ f = \frac{1}{2RC} \]

where:
- \( f \) = frequency in cycles per second
- \( R \) = resistance of the grid circuit in ohms
- \( C \) = capacitance of the coupling capacitor in farads

A convenient method of varying the frequency is by using a dual potentiometer as the grid resistors.

In the symmetrical multivibrator, i.e., one in which equal values of grid resistors and coupling condensers are employed, the output waveforms will resemble Fig. 1a of Part I.
Audio Frequency Distortion Measurements

Part 1, Methods of Measurement

This is Part 1 of a series of four which will deal with audio frequency distortion measurements. Part 2 will give details of a simple practical instrument designed to measure distortion in audio amplifiers.

The theoretical quality of an audio amplifier is related to the amount of distortion prevalent in the amplifier. If, as in true Class A operation, the output plate current waveform of the amplifier should duplicate the waveform of the grid voltage input, such being the case, the amplifier has a certain percentage of harmonic distortion which, if excessive, detracts from the audio quality and becomes annoying to the listener.

Types of Distortion

There are three types of distortion found in an amplifier: (1) amplitude distortion, (2) frequency distortion, and (3) phase shift. In amplitude distortion, the fundamental plus harmonics are observed in the output. Frequency distortion is caused by the amplifier's inability to amplify all frequencies equally. Phase shift is present when the amplifier has different delays for all frequencies. The amount of distortion increases as the tube is operated outside of the linear portion of the tube characteristic curve, as shown in Fig. 1.

In addition to harmonic distortion, there is intermodulation distortion in audio amplifiers. Both are caused by non-linearity in the amplifier. Intermodulation results in the production of frequencies equal to the sum and differences of a low and high frequency (and harmonics). The intermodulation products of fundamental frequencies $f_1$ and $f_2$ are as follows:

- $f_1 + f_2$ and $f_0 - f_2$
- $2f_2 + f_0 + 2f_1 - f_2$
- $f_2 + 2f_0$ and $f_1 - 2f_2$, etc.

The intermodulation products do not resemble the original tones in the input.

Intermodulation distortion measurements more closely correspond to the non-linear distortion detected by the average radio listener than does a measurement of total harmonic distortion. It is interesting to note that intermodulation distortion can be observed even after no harmonic distortion is measurable.

The percentage of total harmonic distortion, expressed as the distortion factor, is equal to

During the positive portion, however, the grid will draw current, which is limited by the series grid resistor. Under these conditions, it is impossible for the grid voltage to rise more than just a few volts positive, no matter how high the input signal goes. Thus, the plate current is also limited. The output will be a square wave. By properly choosing the values of the circuit elements, the limiting action may be made to occur over a wide range of signal amplitudes. Thus, it is possible to convert any complex waveform into a flat-topped waveform.

The circuit of Fig. 8, variously termed a diode clipper, clamping, or limiter, performs a similar function. Any waveform applied at the input terminals will be clipped at a voltage level determined by the bias voltage in series with each diode. This circuit provides a simple method of converting a sine wave to a square wave. Diodes of either the vacuum tube type or the crystal type may be employed.
and is measured on the distortion meter. The distorted wave shape can be represented by the Fourier series, and the relative values of the terms of the series indicate the amplitudes of the harmonics in the composite.

The harmonic content of the signal includes all of the components which are higher in frequency than the fundamental. Component frequencies which are lower than the fundamental, such as noise from the power supply, are not usually measured. Total harmonic distortion measurements are most frequently made at 400 or 100 cycles per second. Even though this is the standard practice, additional distortion will usually be present at the lower frequencies. The Federal Communications Commission recommends a measurement of harmonics in audio equipment at frequencies of 30, 50, 100, 400, 1000, 5000, 7500, and 15,000 cps.

**Distortion Meter**

Audio frequency distortion measurements can be made by using distortion meters, harmonic wave analyzers, and intermodulation analyzers. The distortion meter will be discussed first.

A distortion meter gives the percentage of total harmonic content and does not tell us how much of each harmonic is present in the output. A block diagram of the meter is shown in Fig. 2. Basically, it consists of a null bridge and a wide band wave meter. The null bridge is tuned to the fundamentals such as 400 cycles per second and the bridge is balanced at this frequency to eliminate harmonics of the fundamental. The vacuum tube voltmeter will then measure only the amplitude of the harmonics. For accurate measurements, the oscillator generating the fundamental test frequency should be completely free of harmonics. In addition, the VTVM should represent the RMS voltage as truly as possible. This can be insured by operating the VTVM in a manner such that the square root of the plate current versus grid voltage is a linear function.

The distortion meter does not indicate which frequencies are present in the complex-distorted wave, and the relative amplitude of each. In addition, certain random noises may be very disturbing to the listener and yet show only a small indication on the meter. Therefore, as in many test measurements, the operator must show sufficient skill to translate the results obtained with the meter into useful data.

A typical commercial distortion meter has a frequency range from 50 to 15,000 cps. The distortion percentage is read directly from a meter with calibrated full-scale deflections of 0.3%, 1%, 3%, 10%, and 30% distortion. A grid vacuum tube voltmeter is used for measuring the percentage of total harmonic distortion. The scale is also calibrated in decibels. 6, 100, 000, 000, and 0000 decibels balanced bridge input circuits are provided. Distortion measurements are made on this instrument with an accuracy of approximately 5%. A distortion-free sine wave oscillator should be used with the meter. Otherwise a residual reading will be made which represents the oscillator distortion rather than that of the amplifier or other audio device being tested. There should be no distortion even at the very low audio frequencies.

**Harmonic Wave Analyzer**

Unlike the distortion meter, the harmonic wave analyzer is a practical method of measuring distortion and indicating separate components. The wave-analyzer tells the operator which frequencies other than the fundamental are in the complex waveform and also gives the amplitude of each harmonic.

Since the analyzer must determine the fundamental frequency and all of the harmonics, it is necessary that the instrument be capable of tuning to each of these frequencies and of measuring the amplitude of each. The analyzer is really nothing more than a highly selective vacuum tube voltmeter, and is similar to the conventional superheterodyne receiver; except that the intermediate frequency is much higher than the input audio signal under observation. The wave analyzer has a very narrow bandwidth, otherwise measurements of harmonic components at the very low audio frequencies would be impossible.

The most commonly known wave analyzer is the heterodyne type. A block diagram is shown in Fig. 3 and is representative of a commercial analyzer. The incoming audio signal is heterodyned with the frequency from a variable frequency oscillator and the resultant frequency is amplified by the IF amplifier and read on a vacuum tube voltmeter. When the difference between the oscillator and the input signal frequency is 50kc, the signal will be tuned to the IF amplifier and the amplitude can be measured on the VTVM. The three-crystal filter
incorporated in the IF amplifier assures a very high selectivity. A resonance curve of the crystal filter is shown in Fig. 4. The bolideyne oscillator covers a frequency range of 34,000 to 48,980 cycles per second but the dial is calibrated from 0 to 10,000 cycles. Assume that the incoming signal is 500 c.p.s. This would correspond to an oscillator frequency of 49,500 since 49,500 plus 500 c.p.s. equals 50 cycles. A difference frequency L-3 (49,100—500) cannot be amplified. The bandwidth is only 4 cycles and harmonics can be measured easily at the lowest audio frequencies. The input impedance is one Megohm, which is sufficiently high to make loading effects negligible. The VTVM is directly calibrated in volts and decibels and a 5% voltage accuracy is obtained on all ranges from 250 microvolts to 300 volts full scale. The frequency calibration is accurate to ± (2% — 1 cycle).

Another commercial wave-analyzer has the feature of variable selectivity for rapid analysis of the complex wave. Where the harmonics are spaced far apart, the bandwidth may be increased, thus making it easier to make measurements. Where the harmonics are closely spaced, as at the very low frequencies, the instrument may be made more selective, to separate harmonics 20 cycles apart. A resonance curve of this analyzer is shown in Fig. 4.

The operation of a wave analyzer involves first tuning the oscillator dial to the fundamental and then adjusting the attenuator until the meter reads full scale. Then the harmonics are found by changing the oscillator frequency dial and recording the amplitude of each.

In addition to the use for measuring the distortion in an amplifier, the wave-analyzer can be used for measuring distortion in oscillators, transmitters and telephone systems. It can also be used to determine the harmonics in power machinery and to analyze noise characteristics.

The Intermodulation Analyzer

A block diagram of a typical commercial intermodulation distortion meter is shown in Fig. 6. The amount of distortion is maximum at the highest and lowest transmitted frequencies. However, this discussion will be concerned mainly with intermodulation distortion measurements at the very low audio frequencies. At low frequencies, maximum power output from a tube is not realized because of impedance changes in transformers and reactances. The power output is similarly reduced at the higher frequencies because of increased leakage and distributed capacity.

The operation, two frequencies shown on the diagram (Fig. 6) as 100 and 7000 c.p.s. are transmitted in the mixer. The purpose of the 7000 c.p.s. signal is to act as a carrier for the low frequency components. These two frequencies are commonly used, but a lower frequency ratio must be used if the amplifier under observation has insufficient bandwidth. For best sensitivity, the amplitude of the lowest frequency should be 12 db above the higher frequency—a voltage ratio of 4 to 1. The output of the mixer is fed to the amplifier under test and its harmonics. The resultant signal, which is 7000 c.p.s., modulated by 100 c.p.s., is amplified and demodulated by the oscillator. It is then fed to a low pass filter to eliminate 7000 c.p.s., and the output is fed to a VTVM where the intermodulation products are present and the period of intermodulation distortion is read directly from the meter.

There is no direct relationship between the percentage of total harmonic content and the percentage of intermodulation distortion. With a 12 db ratio for the above frequencies, some authorities claim the percent intermodulation distortion is
equivalent to:

\[ \{h_1, h_2, \ldots, h_n\} X (s) \]

where \( h_1, h_2, \ldots, h_n \) are the harmonics and is the order of harmonics. As an example, 10% intermodulation distortion is often equivalent to about 2.5% total harmonic distortion. Since there are no definite standards for these measurements, any figure of intermodulation distortion must be accompanied by a statement of test conditions.

It is hoped that this brief discussion of audio frequency distortion measurements will be helpful in clarifying the general subject distortion measurements. Part 3 will be especially helpful to those who wish to construct a simple meter for rapid measurement of the percentage of total harmonic distortion and identification of harmonic wave components.

Audio Frequency Distortion Measurements
Part 2, A Practical Distortion Analyzer

Part 1 of this series contained a discussion of the nature of audio frequency distortion and a survey of the methods employed in making quantitative distortion measurements on audio equipment. The present discussion details the design and construction of a simple and practical distortion analyzer which is a very useful adjunct to any amplifier servis shop or audio high fidelity equipment bench. The instrument is compact, easy to adjust and use, and costs little to build. Yet, the results obtained are sufficiently accurate to permit evaluation of the performance of most audio equipment and observation of the results of even minor design changes.

As was pointed out in Part 1, the simplest form of distortion meter employs a null bridge to suppress the fundamental test frequency being amplified under test and a vacuum tube voltmeter to read the amplitude of any signals which pass unattenuated through the null bridge. If the signal input to the amplifier is a pure sine wave of frequency equal to the null frequency of the bridge, the only signal indicated by the voltmeter will be the harmonics introduced by distortion in the amplifier being tested. If the response of the voltmeter is linear, it is easy to express the total harmonic content thus indicated as a percentage of the amplifier output.

The Distortion Analyzer

The major shortcoming of the null bridge type of distortion meter, as it is usually employed, lies in its inability to identify the order of the harmonic content indicated. It reads total percentage of distortion and thus may only be classed as a distortion meter. To be considered a distortion analyzer, the instrument should be capable of identifying each harmonic component present and indicating their relative amplitudes. Commercial distortion analyzers which accomplish this are both complicated and costly. However, a simple system is available which is not appreciably more complicated than the common null bridge distortion meter, but is capable of considerably better results. Its use is predicated upon the availability of a second audio oscillator.

The circuit of a typical null bridge distortion meter is shown in Fig. 1. The components L, C1, C2 and R1 constitute the null bridge network which suppresses the frequency at which L and the series combination of C1 and C2 are resonant, as given by:

\[ f = \frac{1}{2\pi \sqrt{LC}} \]

Where \( f \) is the null frequency in cycles per second. C1+eV1+eV2 is the capacitance of each arm of the bridge.

The circuit configuration will be recognized as the "bridge" type of network. The resistance (R1) is used to adjust the null reading to minimum. If the circuit constants are chosen properly, and distributed capacitance is minimized, virtually zero transmission will occur at the null frequency. If the null circuit "Q" is high, the null will be very sharp and nearby frequencies will be very slightly affected. The voltmeter is used to measure both total amplifier output and harmonic output by shorting out the bridge circuit with the switch (S1) during the former measurement. A vacuum tube voltmeter may be employed, as illustrated in
Fig. 1, or a simple crystal slide voltmeter may be used with only a slight sacrifice in accuracy.

To convert the distortion meter of Fig. 1 to a distortion analyzer, the modifications shown in Fig. 2 are made. An audio transformer is added to permit the insertion of a sine wave signal from a second audio oscillator. This signal is used to identify individual harmonic components present in the bridge output to the best method. To accomplish this, the second oscillator is swept through the frequency range containing the harmonics of the fundamental test signal. Fig. 3 is a block diagram of the complete test set up. Near the frequency of each harmonic present, a "beat" will be observed in the distortion meter reading. The amplitudes of the beats are indicative of the relative magnitude of each harmonic component identified. Thus, a quantitative indication of harmonic content, as well as total harmonic percentage, is obtained.

As an example, suppose that the test frequency is 400 cycles and the distortion meter indicates a total harmonic distortion of 10% before the introduction of the "search" oscillator. If there is both second and third harmonic distortion, an amplitude beat will be observed when the second oscillator is swept through 800 and 1200 cycles. If the second harmonic predominates, the beat at 800 cycles will be greater than the one at 1200 cps, in the same proportion. Knowing the total harmonic distortion, it is easy to evaluate the percentage of each harmonic component. The search oscillator frequency should be adjusted close enough to the harmonic frequency to give nearly zero beat, so that the meter needle can follow. The oscillator used for searching should be reactive free of hum and harmonic output.

Construction Details

The practical circuit diagram of the distortion analyzer is given in Fig. 4. Since no vacuum tubes or power sources are required for its operation, the unit may be assembled in very compact form. A cracked-light metal cabinet, measuring 6" x 5" x 4," affords more than sufficient space to mount all components. No chassis is used; all parts are mounted on the front panel except the chokes (L) and the audio transformer (C) which are supported by a sheet-metal shelf fastened to the back of the removable front panel by means of the shaft bushings of R1, R2, and S1 (Fig. 4). The dimensions of this shelf and the approximate locations of the parts mounted on it are shown in Fig. 5. The suggested front panel layout is shown in Fig. 6.

To assure maximum versatility, three null bridge frequencies: 400, 1000 and 3000 cycles, are provided. These frequencies are selected by substituting the proper capacitance values for C1 and C2. Capacitive switching is done with a two-circuit, three-position switch. If additional or alternative test frequencies are required, the necessary capacitor values may be computed from:

\[ \text{(2)} \]

Where: \( f \) is the desired null frequency.

For most routine amplifier testing, the three frequencies for which values are given in Fig. 4 will be sufficient.

For effective fundamental frequency rejection with low harmonic frequency attenuation, the "O" of the null bridge components must be high. Best quality components should be used for the resonant circuit comprising C1, C2 and L1. The choice of the choke is important since the resistance as well as the inductance of this unit is critical. The resistance of the choke will adversely affect the "O" of the null circuit if too high.

Some selection of capacitors may be necessary to arrive at any given test frequency, although for most practical purposes it is not necessary to make the distortion at exactly the frequencies specified.

The null remote (R1) is a variable one megohm potentiometer mounted
on the front panel and provided with a small knob. This control is used to adjust the null response to minimum at each of the test frequencies. The setting of R2 usually remains fixed for any given frequency.

The crystal diode voltmeter uses a 2N234 or any of the germanium crystals as a rectifier. It gives a response that is approximately linear with input voltage if a high sensitivity meter is used. A 0.1-100 microampere meter is ideal, since the scale calibration can be used to indicate distortion percentage directly. Otherwise, any meter requiring less than about 250 microamperes for full scale deflection may be employed. Above this current, the average crystal diode characteristic departs markedly from linearity.

Two meter ranges are provided to allow more accurate reading of distortion percentages. These ranges, 0-100% and 0-10%, are selected by switching meter multiple resistors R3 and R4 by means of a toggle switch (S1). The amplifier resistor for the 0-10% scale is selected to give full scale deflection at 1/10th the rms input voltage required to give full scale reading on the 0-100% range.

The audio transformer (T) may be almost any unit of good quality which the experimenter might have available. The characteristics are not critical, since this transformer is used merely to introduce a small audio voltage from the search oscillator into the voltmeter circuit. A good 3:1 interstage audio transformer will usually be found satisfactory.

The audio input cable to the bridge circuit, as well as the external lead to the search oscillator, are not through holes in the lower frame of the meter. All cabinet and wiring is permanently to the circuit. These leads are of standard shielded audio cable and are fitted with alligator clips at the input ends. The cabinet holes should be fitted with rubber grommets.

Using the Distortion Analyzer

The use of the instrument is relatively simple. After the construction has been completed, the operation of the null bridge circuit is tested at each of the test frequencies. To do this, the bridge input cable is connected directly to the output terminals of the test oscillator. With the toggle switch, S1 in the "Null" position and the test oscillator and frequency selector switch set at the proper test frequency, the output of the test oscillator and the gain control (R2) are adjusted to give full scale deflection of the deflection meter. Then, with S1 thrown to the "Read" position, the meter reading should drop to a very low value. To minimize the reading, the null resistor (R1) and the test oscillator frequency must be varied simultaneously. If the null bridge is functioning properly, the indicated null at some frequency near the desired test frequency will be quite sharp and the meter reading will be very much zero.

If an incomplete null is obtained, the bridge components are faulty or the test oscillator has some harmonic output which is being indicated on the meter. The nature of this residual reading can be readily determined by the use of the search oscillator. With the distortion meter operating as above in the "Read" position, the search oscillator is connected to the audio transformer input leads and enough search signal is injected to almost double the residual reading on the meter. The frequency range of the test signal and its harmonics is then explored by varying the frequency dial of the search oscillator slowly. If there is a large beat frequency of the meter pointer at the fundamental frequency and little or no multiples, the residual reading is caused by imperfect bridge balance. If the converter is true, the harmonic content of the test oscillator is to blame for the incomplete null. The harmonic output of the test oscillator should be carefully recorded so that it can be discounted when actual amplifier tests are being made.

In using the bridge to analyze the distortion introduced by an amplifier, the procedure followed is the same as that used above for determining the distortion content of the oscillator except that the amplifier is introduced between the null oscillator and the bridge, as shown in Fig. 3. The bridge input leads are connected directly across the speaker voice coil or other normal amplifier load. The gain of the amplifier is set to the value at which it is desired to determine the distortion. The null reading is then compared as above and expressed as a percentage of the full scale reading of the meter minus the residual reading, the total distortion percentage introduced by the amplifier. The harmonic components may then be individually identified by the use of the search oscillator. Each beat noted indicated a component of that frequency (read from the search oscillator) and relative magnitude present in the output of the amplifier.
SECTION V
METERS AND MEASUREMENTS

The Direct-Current Meter

ALTHOUGH the d.c. meter is a standard tool around the laboratory, service bench or "helm shank," its usefulness may be greatly enhanced by a better understanding of the principles underlying its construction and applications. Despite the fact that the judicious use of electrical instruments is an unfolding hallmark of the skilled electronics technician, there is a tendency on the part of many to accept the meter at its face value without ever gaining an intimate knowledge of its internal functioning. Actually, a complete familiarity with the capabilities and limitations of the d.c. meter can be gained only through a study of its electrical and mechanical characteristics. We will presently discuss these characteristics and point out certain precautions to be observed in the use of such measuring instruments. Because the moving-coil permanent-magnet type is the most commonly used, being used to measure current, voltage and resistance with different auxiliary circuitry, the present discussion will be restricted to this type.

The D'Arsonval Movement

The fundamental principle of all general types of electrical meters is the same: electrical quantity to be measured is converted into a mechanical motion, which is proportional in terms of that electrical quantity by means of a scale and pointer. In the D'Arsonval type, direct current flowing in the turns of a coil suspended in a steady magnetic field produces an electromotive force which rotates the armature against the counter torque of a hair—by an amount proportional to the current flowing. A light attached to the armature indicates the rotation of the coil, and therefore the current value, on a semi-circular calibrated scale. Figure 1 illustrates the usual form of this arrangement. The current-carrying coil is wound on a light-weight frame or armature which, in turn, is supported between sapphire-jewelled pivot bearings which allow it to rotate freely. The electrical connections to the coil are made through spiral hair-springs at each end of the armature. These fine alloy springs perform several vital functions. Besides providing the current-carrying path between the armature and the stationary parts of the meter, they provide the counterforce against which the meter torque or rotational force acts, as well as supplying the reactive force which returns the pointer to zero when current ceases to flow. The coil thus mounted is immersed in a strong magnetic field which is usually provided by a permanent magnet. The stability and permanence of this magnetic field are of importance, as well as the uniformity of its magnetic field produced between its poles. The pole tips are usually semicircular in shape to fit closely around the moving coil. The uniformity of field is greatly improved by the use of a cylindrical core of soft iron mounted in the center of the armature so that the moving coil revolves around it. The indicating pointer is affixed to the armature at one end and a system of small adjustable counterweights is used on the tail-piece and inner arm of the pointer to balance the complete armature assembly. The angular movement of the moving coil assembly is restricted by a set of cushioned stops.

The completed assembly is extremely delicate and precise. It is interesting to note that most of the components serve several purposes. For instance, the armature frame not only provides the form upon which the current-carrying coil is supported, but is also a closed-loop conductor in which eddy currents are induced which oppose the motion of the armature and so provide damping of the meter movement. Excessive over-swing or oscillation of the pointer is thus avoided.

The Current Meter

Essentially, the D'Arsonval meter is a current measuring device. The flow of direct current through the moving coil sets up a magnetic field around the coil which interacts with the fixed field produced by the permanent magnet to cause rotation of the coil. The turning torque developed is proportional to the strength of the permanent magnet. The number of turns in the coil, and the amount of current flowing in the coil, the pointer deflection which results is determined by the strength of counter-torque of the spiral springs. At any given meter deflection, the torque produced

FIG. 1

ESSENTIAL PARTS OF D.C. METER
the internal resistance of the meter is connected in parallel with it, the current will divide equally between the two paths and hence twice as much current will be required to give full-scale deflection of the meter. If a shunt is chosen which has one-fourth the resistance of the meter coil, the currents through the parallel resistances divide in the ratio of 4 to 1, and once only one-fourth of the total current flows through the meter, its full-scale indication is multiplied by a factor of 4. Figure 2 shows the connection of a shunt to a direct-current meter and the equation commonly used to determine the shunt resistance required to extend the scale by a factor N. The internal resistance of the meter may be determined from the published characteristics of that type, or by measurement. In most instruments it is usual to select shunts which multiply the scale calibration by multiples of ten for digits in reading.

The D.C. Voltmeter

The same basic movement which is used to measure direct current is also employed in voltmeters. In this case, resistance is added in series with the meter in the manner shown in Fig. 3. Such external multiplier resistances may be used with a high-sensitivity millivoltmeter or microvoltmeter to measure voltages ranging from millivolts to kilovolts. The meter is still performing its normal function as a current measuring instrument, but full-scale deflection of the meter which an unknown voltage causes to flow in a known resistance. The voltage is therefore determined by Ohm's Law: (E = IR) and the actual scale may be calibrated directly in terms of voltage. Meters for voltmeter applications are calibrated according to "ohms-per-volt" ratings, i.e., the number of ohms which must be contained in the voltmeter circuit for each volt which the meter is to indicate. Therefore, to limit a voltmeter using a non-sensitivity bases movement to full scale deflection when 10 volts is impressed, the total resistance of the circuit must equal 30,000 ohms, by Ohm's Law. A total of 15,000 ohms would be required for 15 volts full scale, etc. Thus a 300-ohm meter with multipliers full scale is rated at "1000 ohms-per-volt." The same meter can be used to read 500 volts full scale by using a 500,000-ohm multiplier in series with it. In each case, the output required multiplier resistance is very large compared with the internal meter resistance, the latter is usually ignored since the error introduced is much less than the reading accuracy of the meter. However, if were desired to make a 1000 ohms-per-volt meter read 50 volatile scale, it would be necessary to include the meter resistance in the total value of 1000 ohms required. If the internal resistance of the meter is 100 ohms, the correct value of the multiplier would be 900 ohms since a 10% error would be introduced if the meter resistance were neglected.

Since the voltmeter is always connected across the voltage drop being measured, it is important to use an instrument having a total resistance which is large compared to the circuits to which it is connected. Otherwise, various interferences result since a low resistance meter "loads" the circuit being measured so that the voltage drops indefinitely are not those which exist in the unmeasured circuit. A simplified example of such misuse of the voltmeter is illustrated in Fig. 4. To reduce such errors, basic meters having full-scale sensitivities of 5 microampere (1000 ohms volts) or 10 microampere (10,000 ohms volts) are used in high-quality voltmeters.

The Ohmmeter

Just as the D'Arsonval current meter is used to determine voltage when the current and resistance are known, it may be used equally well to read resistance by indicating the current which flows when a known voltage is impressed across an unknown value of resistance.

Such an instrument, calibrated directly in ohms, is called an "ohmmeter" and is widely used in a variety of circuits, types of which Fig. 6 is a typical example. In this circuit, a
battery or other source of voltage is provided which is capable of producing a full-scale deflection on the meter when the test terminals (A and B in Fig. 5) are shorted. Variations in battery voltage and other circuit constants are compensated for by adjustment of a resistor (R2). If an unknown resistance is inserted between the test terminals, the meter deflection will be reduced proportionately. The meter scale can therefore be calibrated directly in terms of the external resistance required to limit the meter current to that value. When the unknown resistance is equal to the internal resistance of the chintz meter, the meter will read half-scale. The formula used for the calibration of this simple ohmmeter type is also shown in Fig. 5. For the measurement of extremely low or high values of resistance, more complex ohmmeter circuits are employed.

Meters Accuracy

Direct current meters are supplied in many degrees of accuracy according to the requirements of the application. Such applications vary extremely from meters for use as primary laboratory standards having accuracies of 1 in 100 percent to mere indicators of the presence or absence of electricity.

Meters rated at better than 1% accuracy fall into the "precision laboratory" category and should be used only in protected, "well-behaved" circuits requiring such high accuracy. They are usually of the "portable" type which are used with the needle in a horizontal position for greater accuracy and have microscales to reduce parallax errors in reading.

In the accuracy range below 1% are the great majority of "general utility" or "panel" meters which are the "work horses" of the electrical instrument family. They are usually mounted in test equipment panels and switchboards in a vertical position. The average accuracy of this class of meters is about 2%.

The accuracy rating of all d.c. meter types is usually given in terms of the percentage of full-scale reading to which the meter is guaranteed. A single range meter reading 100 volts full scale and rated at 1% accuracy would thus read within 1 volt of the correct value at any deflection. At 10 volts this meter could, therefore, be in error by as much as 1 volt, or 10%.

Good engineering practice dictates that meters be used at a minimum of one-third full-scale deflection for this and other reasons.

Factors Affecting Meter Accuracy

The manufactures nominal accuracy rating does not insure accurate results from a meter in the hands of an inexperienced technician or an instrument which has been subjected to abuse. The following tabulates some of the mechanical and operational factors which may cause large errors in the reading of d.c. meters of the D'Arsonval type:

(a) Stray Magnetic Field Errors. Since the deflection of the meter depends on the strength of the permanent magnet, errors may be introduced by both magnetic fields from other meters, current carrying conductors, magnets and other ferromagnetic materials. Excessive meters are usually provided with adequate magnetic shielding. Some errors are also caused by mounting small meters in heavy steel panels. Meters especially calibrated for such mounting are usually so marked.

(b) Overload Errors. Permanent damage or burn-out may be caused by repeated or heavy overloads of the meter movement. Excessive current through moving-coil type causes heating of the coil and springs. Heating of the latter results in "arresting" or loss of spring tension which impairs accuracy. Overloads also cause needle "banging" which may damage pointer or pivots.

(c) Stick Movement Errors. The meter movement may be prevented from moving freely by several mechanical defects. Chief among these is chipped jewels or damaged pivot due to rough handling. Sticking may be evident in the failure of the meter to reproduce a known reading when approached from values above and below the known value. Light tapping of the meter case is frequently resorted to as a cure. Meter sticking is also caused by small magnetic particles settling on the coil or core of the magnet of a meter which is removed from its case and left unprotected.

High-Resistance Non-Electronic D.C. Voltmeters

A VOLTMETER used to check high-resistance electronic circuits must have high internal resistance, in order to minimize circuit loading. The a c. vacuum-tube voltmeter meets this requirement and has been obtainable in the service category for about 15 years. Most of these electronic voltmeters have input re-
The non-electronic d. c. voltmeter offers the advantages of complete portability, simplicity of operation, freedom from drift and zero adjusting, and the ability to operate without any sort of power supply. These features often are definitely required in field testing and are desirable also in the laboratory when removal from power line and batteries is a requisite. But the common non-electronic voltmeter has relatively low input resistance. It is of interest to note that a non-electronic d. c. voltmeter having high input resistance can be obtained with a sensitive d. c. microammeter and high-resistance multiplier. The full-scale deflection of the microammeter must be somewhat lower than is common in meters ordinarily used as conventional voltmeters. Thus a 0-10 d. c. microammeter with a 25-meg. ohm series resistor provides a 250 d. c. voltmeter. Note that the input resistance in this case is higher than that of the conventional d. c. vacuum-tube voltmeter. The instrument sensitivity is 100,000 ohms per volt.

Ultra-high-resistance, non-electro nic d. c. voltmeters of this type are entirely practical. Multiplier resistors may be switched in the instrument circuit, in the conventional manner, to change ranges. D. C. microammeters are available with full-scale deflections of 3, 5, 10, 15, 20, and 30 microamperes.

<table>
<thead>
<tr>
<th>SCALE</th>
<th>METER RESISTANCE, $r_n$ (ohms)</th>
<th>VOLTMETER SENSITIVITY (ohms per volt)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0-2</td>
<td>10,000</td>
<td>500,000</td>
</tr>
<tr>
<td>0-5</td>
<td>5000</td>
<td>200,000</td>
</tr>
<tr>
<td>0-10</td>
<td>2000-4000</td>
<td>100,000</td>
</tr>
<tr>
<td>0-20</td>
<td>5000</td>
<td>66,466</td>
</tr>
<tr>
<td>0-30</td>
<td>1520</td>
<td>56,000</td>
</tr>
<tr>
<td></td>
<td>1520</td>
<td>33,333</td>
</tr>
</tbody>
</table>

**TABLE OF MICROAMMETER DATA**

**FIGURE 2**

U. R. R. Voltmeter Circuits

Figure 1 shows the circuit of a multirange ultra-high-resistance d. c. voltmeter having 2000, 400, 200, 100, 50, 30, and 15 microamperes full-scale deflections. The voltmeter is of the full-scale deflection type and contains all the elements required to complete the circuit. The 0-10 d. c. microammeter is employed. The voltage ranges provided are 0-0.1, 0-0.5, 0-5, 0-10, 0-200, 0-500, and 0-3000 volts.

The internal resistance (of) of high-sensitivity microammeters is comparatively high. In the 0-10 microammeter, for example, this value lies between 2000 and 4000 ohms, depending upon manufacturer and model number. By heat accuracy, the meter resistance, $r_n$, must be subtracted from the calculated multiplier resistance, $R$, whenever $R/n$ is $100$ or less. Thus in Figure 1, the calculated value of the 0-100 microammeter would be 100,000 ohms. But the meter internal resistance is 2000 ohms, so we must subtract the meter resistance, giving the accurate multiplier value of 98,000 ohms. On each other voltage range, $R/n$ is higher than $100$, so the calculated multiplier resistance values are used.

Figure 2 is a table showing pertinent data for commonly available sensitive d. c. microammeters with full-scale deflections between 5 and 50 microamperes. Displayed in the chart are the internal resistance values and the voltmeter sensitivities (in ohms per volt) which the meters will provide in voltmeter circuits. All except the 2 microampere model are panel-type instruments. The 2 microammeter is a portable case-type, but can be mounted on the panel of an assembled u. h. r. voltmeter.

Figure 2 shows a useful variation of the u. h. r. voltmeter circuit. Here, a center-zero type of microammeter is employed. The right half of the scale is graduated from zero up to the maximum positive voltage and the left half from zero down to the maximum negative voltage of the same value. When a test voltage is applied so that the upper input terminal is positive, the meter is de-
These relationships point up the necessity for using the lowest-range microammeter that can be afforded for this application.

Pointers on the Voltmeter

The combination of extremely sensitive microammeter and very high multiplier resistances is not common-place. Certain precautions are necessary in the construction of voltmeters employing this combination, which do not arise in connection with microammeters up to 20,000 ohm-per-volt sensitivity. These precautions are outlined below.

1. Special instrument-type resistors are required beyond 50 microamps. Be sure that the outer surfaces of these components are cleaned carefully after soldering into place. Try not to touch the surfaces with the bare fingers, to prevent depositing grease or moisture onto the resistors. After installation, wash the exterior portions of the resistors with carbon tetrachloride or any other solvent recommended by the resistor manufacturer. If the resistors have been coated with a special high-insulation wax, DO NOT TOUCH THE BODIES, and do not use solvents which might dissolve the wax.

2. Use the minimum of heat in soldering the resistors in place. Provide a protective heat sink by holding the joints with flat-nose pliers while making sure to grip the leads with the pliers until the soldered joint has cooled completely.

3. The range switch must be provided with excellent insulation to prevent leakage paths. When a ceramic-type switch is used, this will handle the ceramic insulation any more than necessary during assembly of the voltmeter. After installation, wash the ceramic with carbon tetrachloride or similar solvent to remove any oil contamination.

4. The input terminals of the voltmeter must be insulated from the potentiometer by ammeter washers or inserts of high-quality dielectric material, such as polytetrafluoro.

5. If a metal panel is used, be sure to specify metal-panel operation when ordering the microammeter.

6. Make an air-tight seal of the instrument case to prevent the entry of dust, grease, and moisture which will send high-resistance leakage paths on the range switch and multiplier resistors.

7. Assemble from the meter manufacturer whether the microammeter can be used in all positions. If the meter is specified to operate in one position only (such as horizontal), be sure to specify the correct position of the meter in all positions.

8. Include a short-circuiting switch for the short-circuiting device if the voltmeter itself when the voltmeter is not in use. Leakage currents from moving parts are susceptible to fields and transient phenomena, and the short circuiting device will prevent them from the voltmeter's output terminals while the shorting switch is closed.

Pointers on the Voltmeter

The special technique is required to measure voltages with the ultra-high-resistance meter. Manipulation of the instrument and its errors are the same as with lower-resistance non-electronic voltmeters. However, it is appropriate to call attention here to a few precautions which should be observed to protect the meter and insure continued accuracy of the instrument.

1. When uncertainty exists as to the approximate level of a voltage, use the lowest-voltage range first. Then, switch down incrementally to each lower range until indication is
obtained in the upper portion of the scale.

2. Switch the voltmeter to its high-est range during idle periods, and also close the microammeter short-circuiting switch.

3. Always close the short-circuiting switch when the instrument is being transported. This provides efficient damping of the meter and will prevent mechanical damage due to shaking and vibration.

4. Do not expose an ultra-high-resistance voltmeter to magnetic fields, moisture, or corrosive fumes for long periods of time.

5. Keep the panel area around the input terminals scrupulously clean of dust, grime, and moisture.

<table>
<thead>
<tr>
<th>Voltage Range</th>
<th>Basic Meter Range</th>
</tr>
</thead>
<tbody>
<tr>
<td>0-2 mV</td>
<td>0-5 mV</td>
</tr>
<tr>
<td>500 K</td>
<td>500 K</td>
</tr>
<tr>
<td>1 mV</td>
<td>1 mV</td>
</tr>
<tr>
<td>1.25 mV</td>
<td>1.25 mV</td>
</tr>
<tr>
<td>2 mV</td>
<td>2 mV</td>
</tr>
<tr>
<td>2.5 mV</td>
<td>2.5 mV</td>
</tr>
<tr>
<td>3 mV</td>
<td>3 mV</td>
</tr>
<tr>
<td>3.75 mV</td>
<td>3.75 mV</td>
</tr>
<tr>
<td>5 mV</td>
<td>5 mV</td>
</tr>
<tr>
<td>7.5 mV</td>
<td>7.5 mV</td>
</tr>
<tr>
<td>10 mV</td>
<td>10 mV</td>
</tr>
<tr>
<td>15 mV</td>
<td>15 mV</td>
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<tr>
<td>20 mV</td>
<td>20 mV</td>
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<tr>
<td>25 mV</td>
<td>25 mV</td>
</tr>
<tr>
<td>30 mV</td>
<td>30 mV</td>
</tr>
<tr>
<td>37.5 mV</td>
<td>37.5 mV</td>
</tr>
<tr>
<td>50 mV</td>
<td>50 mV</td>
</tr>
<tr>
<td>75 mV</td>
<td>75 mV</td>
</tr>
<tr>
<td>100 mV</td>
<td>100 mV</td>
</tr>
<tr>
<td>150 mV</td>
<td>150 mV</td>
</tr>
<tr>
<td>200 mV</td>
<td>200 mV</td>
</tr>
<tr>
<td>250 mV</td>
<td>250 mV</td>
</tr>
<tr>
<td>300 mV</td>
<td>300 mV</td>
</tr>
<tr>
<td>375 mV</td>
<td>375 mV</td>
</tr>
<tr>
<td>500 mV</td>
<td>500 mV</td>
</tr>
<tr>
<td>750 mV</td>
<td>750 mV</td>
</tr>
</tbody>
</table>

**Table of Multiplier Resistance Values**

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**Applications Of The Electrometer**

The electrometer is an invaluable instrument in physics and electrical engineering when its potentialities and peculiarities are understood. It finds use both in research and testing. In one form or another, the electromechanical type of electrometer has been used in experimental physics for many years. Its familiar types include quadrant, bendix, and fiber. Electrometers differ from elec-
FIGURE 1.

CHARACTERISTICS OF ELECTROMETER TUBES

FIGURE 2.

TYPICAL VT ELECTROMETER CIRCUIT

73
Figure 2 shows a typical skeleton circuit of a vacuum-tube electrometer. The circuit is battery-operated. V₁ is the filament battery, V₂ plate battery, and V₃ a bucking battery for zero setting. The “high” input terminal, X₁, is provided with a guard ring. Terminal X₂ may be grounded to X₃ or sealed, as test conditions require. The indicating d. c. microammeter is connected in the “cathode” return circuit in series with resistor R₃ which is kept high in value for maximum degeneration stability and linearity are enhanced by this degeneration. The instrument is set to zero by means of potentiometer X₅ and the bucking battery V₃. Switches S₁ and S₂ disconnect the battery when the electrometer is not in use. V₅ acts as a-safety switch is required, since disabling the filament circuit removes plate current.

A standard radio tube would be unsatisfactory in this circuit, since its input (grid-filament) resistance would be too low for electrometer use. Maximum current amplification demands that input resistance be high. Tube insulation usually is good, but internal charges reach the grid, increasing conductance. Special electrometer tubes have good evaporation and operate at low plate voltage to prevent ionization of whatever residual gas is present. They are operated at low filament voltage and current, and some types are provided with an internal shield grid to isolate the control grid from positive ions from the filament. In some instruments, the second grid is used for the test-voltage input. In the electrometer, the tube is darkened to prevent spurious photoelectric effects, its envelope is washed carefully with a grease-removing solvent, and the outer surface of the envelope...
may be coated with a high-quality insulating material for additional protection against contamination from accidental touching or from the atmosphere. Figure 1 shows the characteristics of some electrometer tubes.

The electrometer usually is provided with several range switches. Range switching is accomplished by changing simultaneously the values of voltages $V_1$, $V_2$, and resistor $R_2$. A portion of $R_2$ is made adjustable for range calibration. The scale of meter $M$ may be calibrated to read directly in volts. Resistor $R_1$ is a current-limiting component, the purpose of which is to limit tube input current when excessive signal voltages are applied.

Excellent insulation is employed for the input terminals, range switch, zero-set potentiometer, and limiting resistor. These precautions are necessary to prevent a low-resistance shunt of the high input resistance of the tube. Input capacitance of the instrument is very low, being less than 10 µfd. In commercial vacuum-tube electrometers. Because of the high input resistance and the consequent long time constant, this capacitance can retain a charge and cause the meter to remain deflected after a test voltage has been removed from the input terminals. For example, a time constant of 7.5 seconds would be obtained with an electrometer having 7.5 microfarads input capacitance and 10 megohms input resistance.

Because of the low electrometer-tube currents, the plate and filament batteries can be expected to give long life, requiring infrequent replacement and causing little concern if the instrument inadvertently is left switched-on during brief idle periods.

Typical electrometer applications are discussed in the following paragraphs.

Current Measurements

An external shunt resistor may be used to convert the electrometer into a current meter in the same manner that such a resistor is used with a
d. c. vacuum-tube voltmeter. (See Fig- 
ure 7A). However, the difference is 
that the electrometer shunt may have a 
high resistance value, often many 
megohms. In this way, the electro-
meter may be converted into a micro-
microammeter. If the scale of the 
electrometer is graduated in volts, the 
unknown current value 1 (in microm- 
amps) = E/R, where E is the 
electrometer deflection in volts, and 
R is the shunt resistor value in 
megohms.

Figure 3(b) shows the electromet-
er and an external shunt conveniently 
may be used to measure the dark 
current of a phototube. The tube is 
supplied with its normal polarizing 
voltage, and the resultant leakage 
current sets up a voltage drop across 
the high-resistance current shunt. 
This drop then deflects the meter 
and is converted into current units 
as explained in the preceding para-
graph.

The tiny leakage current of an in-
ulated sample, especially at low test 
voltages, may be checked with the 
arrangement shown in Figure 3(c). 
This circuit is analogous to the pre-
ceding one, in that a d. c. test volt-
age is applied to the sample in series 
with a high-resistance current shunt. 
The polarizing voltage will be of 
whatever value at which the leakage 
does not occur. This arrangement 
may be employed also for checking 
the leakage current of high-quality 
non-electrolytic capacitors such as 
mercury, ceramic, and first-grade paper 
types.

The electrometer can be used to 
check ion-chamber current in radio-
activity tests in the manner illus-
trated by Figure 3(d). The polariz-
ing voltage is applied in the cor-
rect polarity as the ion chamber in 
series with a high-grade capacitor, 
C, of accurately known capacitance. 
Radiation causes ion charges to be 
stored by capacitor C. The electro-
meter voltage deflection, E, then is 
proportional to the radiation (Q = 
Ck).

Other current measurements appli-
cations of the electrometer include 
checking of (1) vacuum-tube grid 
current, (2) low-order currents in 
dimly-illuminated photocells and 
phototubes, and (3) surface leakage 
on insulators.

Voltage Measurements
D. C. potentials at high resistance 
are checked readily with the elec-
trometer in many circuits in which 
the comparatively high resistance of 
a vacuum-tube voltmeter is entirely 
desirable. When the test voltage 
exceeds the maximum deflection pro-
vided by the electrometer range 
switching, an external high-resistance 
voltage divider (R1, R2) may be used, 
as shown in Figure 4. The unknown 
voltage, E, then will be equal to 
(E/R1, R2)/R2, where E is the elec-
trometer deflection in volts, and 
R1 and R2 (the voltage-divider resistance 
arms) are in ohms or megohms each.

Figure 4(B) illustrates measure-
ment of the open-circuit voltage of 
a d. c. power supply having high in-
ternal resistance, R0. The accurate 
measurement of such a terminal volt-
age (a resistive-limited constant-cur-
rent transistor bias supply is an ex-
ample) would pose a problem if only 
a v. t. voltmeter were available, since 
the internal resistance R0 would form 
a voltage divider with the voltmeter 
input resistance.

The arrangement in Figure 4(c) 
measures the measurement of the 
charged voltage and leakage rate of 
a sample capacitor, C. E1 and R2, if

![Electrometer Diagram](image-url)
required, are chosen in value so that their total resistance is much higher than the leakage resistance of the capacitor. Switch S is thrown first to position A. This connects the capacitor across the polarizing-volt-
age source and charges it. The switch then is thrown to position B, connecting the charged capacitor to the electrometer through the voltage divider. The initial detection of the electrometer shows the charged volt-
age of the capacitor, and the decline of this reading with respect to time indicates the discharge of the capac-
tor. This type of test perhaps is more indicative and valid when the external voltage divider (R₁, R₂) can be omitted. Then capacitor C looks into the very high resistance of the electrometer.

The static potential at a vacuum-
tube electrode in series with a high resistance is measured accurately with the electrometer. The control grid is an example. Figure 4(D)

shows the connections for checking static grid potential across a high value of grid resistance, R₉.

Other applications involving the use of an electrometer to measure potentials include checking of (1) piezoelectric crystal voltage, (2) out-
put of slightly-biased thermocouples, (3) contact potentials, (4) static electr-

city, and (5) physiological poten-
tials in biological and medical re-

search.

Capacitance and Resistance

Measurements

The high input resistance of the electrometer permits determination of capacitance by d.c. methods. Fig-


u m 5(A) is an example. Here, C₀

is a high-grade standard capacitor of accurately-known capacitance and ex-

cellent leakage characteristics. The polarizing voltage, E, also is known accurately. The capacitor shunts the

electrometer input terminals. Capac-
te C₀ is the unknown unit. Capa-

citance is determined in terms of charge division between the standard and unknown. With switch S in position A, the capacitors charge in series. The voltage reading, E, of the electrometer is noted. The unknown capacitance C = (C₀E)/E₀.

High resistance values may be de-
termined from the measured time constant of a circuit containing the resistance (R₁) and a charged capaci-
tor (C₀) of accurately-known capac-

citance, as shown in Figure 5(B). When switch S is closed, capacitor C₀ is charged to the potential of the polarizing voltage, and this value is indicated by the electrometer. When the switch is opened, the capacitor begins to discharge through R₁. The discharge rate then is accurately timed up to the point at which the electrometer voltage deflection has fallen to 37% of its initial value. The

unknown resistance R₁ in (megohms) = t/C₀, where t is the discharge time (in seconds) and C₀ the stand-

ard capacitance in microfarads.

Testing Semiconductor Diodes

T HESE seems to be a growing im-

pression that semiconductor di-

odes can be tested adequately with an ohmmeter. This results from the fact that a 5c ohmmeter will show a difference between the forward and reverse resistances of a diode or rect-

ifier if its leads are swapped back and forth.

The unsuitability of the ohmmeter test as a sole check method lies in the fact that the meter voltage and current often bear no significant rela-
tionship to the rated d.c. parameters of the diode. Thus, a diode may be checked as good at one ohmmeter voltage and still be unsatisfactory at the rated voltage, and vice versa. Al-

so, it very often is difficult to deter-
mine by this method the actual front-
to-back resistance ratio of a diode; since the ohmmeter must be operated on at least two different ranges in order to read the values accurately, and switching the ranges not only changes the applied voltage but also the value of series resistance intro-
duced by the instrument. The conse-
quent variations in applied voltage and load resistance change the oper-
ating point along the diode charac-
teristic curve, separate diodes also have the same effect, and the read-
ings are meaningless unless all fac-
tors and conditions are known.

Another important consideration is that a diode might check satisfac-
tory in a d.c test, ohmmeter-type or otherwise, yet not be suitable for an intended application as an a.c. or r.f. rectifier or demodulator.

Use of the ohmmeter therefore is restricted to the simplest sort of initial test for separating good di-

odes from bad, since all that this instru-

ment indicates reliably is that the component under test is a rectifi-

cer.

Types of Tests

Two types of tests may be applied to semiconductor rectifiers, whether crystal diodes or power rectifiers. These are identified broadly as d.c.
tests and a.c. tests. There can be several categories in each type.

Basically, the d.c. test consists of passing a specified amount of cur-

rent through the diode and checking the resulting d.c. voltage drop across the diode. The test is made separ-
ately, with the proper polarities, for forward and reverse conduction through the diode. Note that this is the opposite of testing tube-type recti-

ifiers, where the procedure is to ap-
ply a specified voltage and measure the resulting current. The d.c. test may be made at a multiplicity of points and a continuous static curve plotted from data taken at these points. A small test consists of data taken at single forward and reverse check points specified by the diode manufac-

turer.

A-c tests fall roughly into two groups: (1) the rectifier test involves applying a given sinusoidal a.c. volt-

age to the diode and measuring the resulting d.c. output current, and (2) the c.m. detector test in which an am-

plitude-modulated sinusoidal voltage is applied to the diode in series with an appropriate load resistance, by-

passed at the carrier frequency, and the resulting modulation-frequency voltage measured across the load re-

cistance.
Other special tests for certain diodes of specific types or for those intended for special-purpose applications will be described later.

As to which type of test is the most suitable, there is general agreement among semiconductors experts that a diode should be tested under the conditions that most closely approximate the intended method of operation. This does not necessarily mean that a diode which is to be used, for example, as a video detector must be tested in the actual tv receiver circuit (although that is not ridiculous). But such a diode should be given an r-f, instead of d-c, test. Likewise, a diode which is to be used as a meter rectifier should be given a low-frequency a-c test. Conversely, a diode which is to be used in a direct-current application, for example as a polarity-sensitive element in series with a relay coil, should be checked at the direct current level which it must operate. Very little significant information could be gained from a d-c test regarding the r-f performance of a diode, nor could the r-f test reveal what might be expected in the way of d-c performance.

D-C Tests

Characteristic Curve. Figure 1 shows the apparatus setup for d-c measurements. The variable-voltage d-c source may be either a battery or a line-operated power supply. The reversing switch allows changeover of the polarity. The voltmeter and current meter also must have reversing switches, although not shown in the schematic.

First, with the anode of the diode under test biased positively for forward conduction, the current is adjusted to several levels within the safe operating range of the diode, starting at zero, by varying the voltage. The current and voltage variations then are recorded. Next, the reversing switches are thrown to bias the anode negatively for reverse (back) conduction and the procedure repeated. The maximum diode voltages, both forward and reverse, and the maximum currents must not exceed the maximum continuous operating voltage given by the diode manufacturer. The current and voltage data so obtained may be used to plot a static curve of the type shown in Figure 2. The forward conduction characteristic extends from 0 to A, and the reverse conduction characteristic from 0 to B. Positive voltage values are low; negative values high.

For small diodes, forward current is expressed in milliamperes and reverse current in microamperes; while for power rectifiers, forward current is in amperes and reverse current in milliamperes.

Figure 2 shows the general shape of a diode static characteristic. Various diode types show different departures, one way or the other, from this curve. In gold-bonded germanium diodes, for example, section OA is steeper than is point-contact type. In germanium power rectifiers, OA is quite steep and OB moderately steep. The selenium rectifier curve has slopes which are intermediate between the preceding types. The silicon junction diode is a special case, its curve showing a characteristic such as Figure 3. The forward current rises sharply at a particular value of positive anode voltage, and the reverse current similarly increases sharply at a particular reverse potential, called the Zener voltage.

The static characteristic curve is helpful when information is desired regarding the behavior of the diode throughout its operating range. Obtaining such a curve by the point-by-point method, however, is laborious.

Single Point D-C Test. In many requirements for a diode acceptance test, a complete characteristic curve is not needed. In these cases, a sin-
gle forward check and single reverse check will suffice.

The same apparatus setup shown in Figure 1, or a similar one, is used. All spot voltages and currents are measured. Most small diodes may be checked at +1 volt and -1 volt. When these diodes are to be used for blocking high reverse bias voltage, they usually are checked at +50 v or -100v, depending upon type.

Observations in D.C. Tests. The main objective of the d.c. test is to determine whether the diode complies with current and voltage specifications. While under test, however, observations may be made of other important features which might render the diode unsatisfactory. These include current or voltage Butler current or voltage drift, heating, and intermittence.

A.C. Tests

Rectification. In this test, an a-c voltage of required frequency and amplitude is applied to the diode or rectifier, and the resulting d.c. output current measured. Figure 4 shows the test setup.

By means of the Variac, the applied a-c voltage, as indicated by the voltmeter, is adjusted to the rated operating voltage of the diode. The output current is read from the current meter. The latter will be a 4 amp, 0.1 ma, selenium, zinc and silicon diodes and an ammeter for power-type germanium, selenium, silicon, copper oxide, and magnesium-copper sulfide rectifiers.

At frequencies other than that of the a-c power line, the Variac must be one designed to handle the operating frequency. At high audio and radio frequencies, a suitable adjustable output signal generator may be substituted for the power line and Variac. The generator output impedance must be low. At high frequencies, a 30 volt milliammeter is needed to measure the applied voltage.

In the rectification test, the d-c output current is noted when the applied voltage is the operating voltage specified by the diode or rectifier manufacturer. The test is repeated at several values of recommended load resistance.

Rectification Efficiency. The ability of a diode to function satisfactorily as a rectifier may be evaluated in terms of the ratio of its d-c output voltage to the applied a-c voltage. This ratio is termed rectification efficiency.

Rectification efficiency often is specified in critical applications of germanium and silicon diodes in communications and instrumentation equipment. Figure 5 shows the apparatus setup for measuring this factor.

The test signal is supplied by a suitable adjustable-output signal generator through an isolating transformer, T. The output winding of this transformer has both low impedance and low d-c resistance with respect to the diode forward resistance.
The type of transformer and its characteristics depend upon the test frequency. This unit will be trim-coded for audio and power-line frequencies, and will be of either air-core, powdered-iron, or ferrite-coated type for radio frequencies.

The input voltage (E₁), indicated by the a-c vacuum-tube voltmeter, is adjusted to the required level, and the d-c output voltage (E₂), developed across the load resistance (R₂), is read from the d-c vacuum-tube voltmeter. The required values of R₁ and C usually are given in the specifications of the diode under test (a common combination is R₁ = 5000 ohms, C = 20 μf, f = 10 Mc). The rectification efficiency (η) equals the d-c voltage (E₂) divided by the peak a-c voltage (E₁). X 1.414.

Expressed as a percentage, this fraction must be multiplied by 100. Thus, 

\[ n = 100E_2 / (1.414E_1) \]

Rectification efficiency varies directly as the a-c voltage amplitude, frequency, and load resistance.

A.M. Detector Test. An important use of small diodes is in detector (demodulator) circuits. In this application, the function of the diode is to separate the modulation component from the carrier component in an amplitude-modulated wave. Figure 6 shows an apparatus setup for checking detector action. The a.m. signal is supplied by a signal generator in which the output and the modulation percentage can be continuously variable. Resistance R₁ is a low resistance and is not required if the output circuit of the signal generator contains a low-impedance return path for the diode under test. The signal level is monitored by the r-f vacuum-tube voltmeter/millivoltmeter. The diode output, developed across the load resistance (R₂), is monitored by the a-c vacuum-tube voltmeter/millivoltmeter.

At a given value of modulation percentage and carrier amplitude, the efficiency of the diode as a detector is proportional to the voltage indicated by the audio meter (E₂ = r-f voltmeter). For a complete evaluation, the test should be repeated at various signal-to-voltage and modulation-percentage levels and at several carrier frequencies. At low signal levels, response of the diode will be observed to be approximately square-law.

The value of load resistance R₂ usually is prescribed in the detector diode specifications. Capacitance C is chosen for effective bypassing of the carrier-frequency content component. If no value is specified for R₂, this resistance should be that value into which the diode will operate when installed in the equipment in which it is to be used.

Television Diode Test. A variation of the detector test setup, for tv diodes, is shown in Figure 7. In this arrangement, a 40 Mc, 70-per- cent-amplitude-modulated signal is applied to the diode under test. This is the arrangement suggested by the Joint Electronic Tube Engineering Council, JETEC (see Sykeson Engineering Information Service, Aug. 1964).

The test procedure is the same as in the a-m detector test described in the preceding Section, except that the a-m frequency and modulation frequency are maintained constant.
Visual Test Method.

It was mentioned earlier that the production of a diode conduction curve, such as Figure 2 is a tedious process requiring the painstaking accumulation of current and voltage values point-by-point. It thus becomes costly to evaluate a large number of diodes by this method.

A visual test method which is dynamic in nature gives a display of the complete curve on an oscilloscope screen. Figure 9(A) shows a circuit for visual display of the diode characteristic. The diode under test is connected in series with an a-c bias supply and a current resistor. This resistance is low with respect to the diode forward resistance and usually is between 1 and 10 ohms.

During the half-cycle of applied 60-cycle voltage when the anode of the diode is positive, high forward current flows through the current resistor. During the negative half-cycle, the anode is biased negatively and small reverse current flows through the resistor. The voltage drop across the current resistor is proportional to the forward and reverse currents and is applied to the vertical input of the oscilloscope. Vertical deflection thus is proportional to current. The transformer output voltage is applied to the horizontal input of the oscilloscope. The horizontal deflection accordingly is proportional to the diode voltage.

The reason for using a d-c oscilloscope is that the direct-coupled nature of its circuitry enables definite establishment and identification of the zero current and voltage point (origin) on the pattern. The horizontal axis may be calibrated in volts (forward and reverse) and the vertical axis in amperes, milliamperes, or microamperes forward and reverse.

Figure 9(B) shows typical patterns displayed by the curve tracer circuit. Pattern (a) is the type normally obtained for a good diode. The negative (reverse) portion of (b) changes its position intermittently, or flutters, indicating instability. This condition occasionally is observed also in the forward (positive) region of the curve. In (c), the reverse portion of the curve is seen to be traced along one route from zero to maximum, but to return along another route from maximum back to zero. This opening-up of the curve is termed hysteresis, a condition which often foretells early failure of the diode.

Checking Recovery Time.

A semiconductor diode has the peculiar property that its reverse current will be relatively high (reverse resistance low) for an instant immediately after applying the reverse voltage if the diode has just been conducting forward current. After this initial high-current transient, the reverse current then decreases gradually (resistance increases) to a value in line with the applied voltage. The interval during which the current settles to its rated level is termed recovery time. Recovery time increases with the level of recent forward current. It is longest in junction diodes than in point-contact types. Short recovery time is particularly desirable in digital computers and similar pulse-type circuits in which diode conduction is switched rapidly from forward to reverse.

The wave pattern in Figure 9(B) illustrates diode reverse-current recovery effect. The diode conducts forward current during the interval from a to b. Instant b corresponds to time t₁. (See Figure 8A). At b,
Improved, Crystal-Type Noise Generator

The satisfactory repair of modern high-gain electronic equipment, such as amplifiers and receivers, as well as the development of this apparatus, often involves some kind of noise level measurement. The technician's attention is being directed increasingly to this subject by the specifications and performance requirements for preamplifiers, preselectors, boosters, complete audio amplifiers, complete video amplifiers, and communications receivers. The normal presence of electrical noise background in high-gain equipment is well-known. Its theoretical aspects have been treated thoroughly elsewhere and will not be discussed here. The technician understands that internal noise arises inherently from current flow and thermal action in tubes, resistors, contacts, etc. and that its presence limits the smallest signal which can be amplified or even handled by a system. While electrical noise possesses a number of distinguishing characteristics, the principal ones are its random nature, the distribution of its energy over an extremely wide band of frequencies, and its characteristic acoustic similarity. Its multicomponent waveform...
contains jagged peaks and amplitude smear.

The residual noise level at the output terminals of an equipment may be measured with an oscilloscope or suitable a-c milliammeter. Generally, the input terminals of the equipment are short-circuited, although some specifications may call for measurements with the input open. The residual output signal can contain, in addition to noise, other components due to hum arising with in the equipment or oscilloscope. Obviously, every effort must be made to separate the noise voltage from any other such components if the noise measurement is to be valid. This is not often easy. It is interesting to note, however, that an occasional specification will lump the total residual small output due to all such factors under the generic heading of noise.

A standard method of noise measurement consists of feeding an adjustable-amplitude noise signal into the input circuit of an equipment under test while monitoring the noise-output power of the equipment, and noting the increase in noise-input amplitude required to double the output-noise power. The input noise then equals the internal noise of the equipment; and from these data, a noise factor may be calculated. If a wattmeter is not available, the noise may be monitored in lieu of noise power, an increase in output voltage of 1.414 times corresponding to doubling the power. Both receivers and amplifiers are checked in this manner.

The noise test signal is derived from a noise generator, of which there are several types. Laboratory versions of the instrument are based upon a specific temperature-limited noise diode tube (example, Sylvania Type T722). When the plate voltage of this tube is adjusted to the level at which the filament emission is saturated (that is, all emitted electrons are collected by the plate), the shot noise generated by the diode has constant amplitude and its energy is distributed over a wide spectrum. At intermediate plate voltages, the diode noise current is proportional to the d-c plate current. Thus, by providing for smooth variation of the diode plate voltage, a noise signal of controllable variable amplitude may be obtained. The a-c noise component of the test signal is coupled, through the noise generator output circuit, to the input of the amplifier or receiver under test.

The circuit of the improved generator is given in Figure 1. In this arrangement, the diode current is adjusted by means of the 40,000-ohm wirewound rheostat, R, The current level is indicated by the milliammeter, M. This noise generator has three ranges, selected by switch S: 0-100 microamperes, 0-1 milliamperes and 0-10 milliamperes.

This arrangement for constant monitoring of the current over a possible range of 5000 to 1 enables reasonably accurate re-setting of the output-noise level. The resistance values of the two noise-shunt resistors, R, and R, are based upon an internal meter resistance of 800 ohms. This is the resistance of the Triplet Model 2017 instrument employed in the writer's prototype. Some variation will be necessary when an individual internal meter resistance differs from this 800-ohm value. In any case, the required shunt resistance (in ohms) equals R, 9 by the 1-ma range, and R, 19 for the 10-ma range. The numerator, R, is the internal meter resistance, in ohms.

The a-c noise component due to random fluctuations of current in the diode is transmitted to the OUTPUT terminals through the 0.001-uf capacitor, C. For the capacitor forms a closed circuit comprising the R1, C, and the OUTPUT terminals.

A non-inductive (good-grade composition or carbon film) resistor, R, bridges the OUTPUT terminals to match the input resistance, or impedance, of the device under test. If the noise generator is used, for example, with a receiver having an input impedance of 300 ohms, R, will be 300 ohms. Resistor R, need not be larger than 1/2 watt. Since many modern input impedances are unencumbered in amplifiers and receivers, the question will arise as to why a switch arrangement has not been employed here to give the noise generator a wide range of output impedances. The reason for not doing this was to avoid the inherent noise-generating properties of such a switching circuit. In fact, in order to keep all internal noise not arising in the diode, and extraneous pickup at a minimum, resistor R, diode
1N21B, and capacitor C must be mounted right at the OUTPUT terminals. This insures the shortest possible leads. When the generator is to operate into a new impedance, Z, a new resistor R1 equal to Z must be fastened directly across the OUTPUT terminals.

When the device under test has a balanced input circuit, as often is the case in television receivers, two series-connected resistors with their junction grounded can be connected, for impedance-matching across the noise generator OUTPUT terminals in the manner shown in Figure 2. The value of each resistor is equal to 1/2 Z, where Z is the input impedance of the device under test.

Whether the generator has balanced output (Figure 2) or unbalanced output (Figure 1), its ground must be connected solidly to the ground of the device under test.

Construction

The simplicity of the noise generator Circuit removes any complication of construction. By using a 6-volt battery (B) of thin, flat construction, like Burgess Type 5000 (4" x 1 1/2" x 3/4"), the entire instrument may be housed in a metal radio utility box slightly larger than a standard meter case.

The insert in Figure 1 shows the shape of the 1N21B diode. The metal tip on the left end of this figure is the cathode terminal which is connected to the junction of rheostat R3 and capacitor C.

The only wiring precaution is to keep all connections between the diode, capacitor, OUTPUT terminals, and R3, as short as possible. To accomplish this, solder the base of the diode directly to the upper OUTPUT terminal, and solder capacitor C from the tip of the diode to the lower OUTPUT terminal. In order to prevent damage to the diode during this operation, hold with metal pliers the metal part of the diode being soldered. This will conduct the heat away from the diode case. Continue to grip with the pliers until the diode is unmistakably cool to the touch.

Operation

Figure 3 shows a typical setup for noise measurements. The following test procedure is recommended: (1) With the noise generator switched off, note the reading of the output meter (A.C. v.t.voltmeter or audio wattmeter) when the amplifier or receiver is in operation and its output controls are set for normal operation. (2) Record this deflection of the output meter, due to the inherent internal noise level of the equipment under test, as the “zero level.” (3) Switch-on the noise meter and increase its output by adjusting rheostat R3, until the output power of the amplifier or receiver (when the output meter is a wattmeter) is doubled, or until the output voltage (when the meter is a voltmeter) is increased 1.414 times. (4) Note the reading of the current meter, M. High current values indicate a high noise level in the device under test, and vice versa. After work is done in the device, a decrease in the initial current reading obtained in a repeated step 4 indicates an improvement in the noise characteristics; an increase in meter M deflection shows a worsening of the noise.

In addition to the comparative type of measurement just described, the noise generator output may be calibrated quantitatively. One method is to plot the noise output voltage (E) at the generator output terminals vs meter M readings for a given output impedance, Z0. Either these voltages might be used in subsequent noise tests, or the noise power levels might be calculated; P = E2/Z0, watts. For either voltage or power, a separate calibration is required for each output impedance, Z0.

The technician will find noise measurements advantageous in checking the performance of all-wave receivers, television receivers, and amplifier systems for sound reproduction. These measurements are particularly revealing when used to determine the efficacy of TV boosters, since the latter devices are known to suffer occasionally from prohibitive noise.

84
SECTION VI
SPECIALIZED APPLICATION AND DEVELOPMENT

Using Standard Time and Frequency Broadcasts

The standard time and frequency transmissions of the National Bureau of Standards radio stations WWV and WWVH provide an invaluable service to laboratories and individual experimenters throughout the world. Extremely precise audio and radio frequency standards, as well as accurate time intervals and radio frequency propagation warnings, are placed at the disposal of anyone having a receiver capable of tuning to one or more of the transmitting frequencies. The proper use of these facilities can be made to greatly supplement the instrumentation of any laboratory. However, the maximum utilization of this valuable "natural resource" depends upon a knowledge of the broadcasting schedules, transmitting frequencies, and suitable methods of comparison.

The standard frequency stations WWV and WWVH are operated by the National Bureau of Standards of the United States Department of Commerce. Station WWV is located at Beltsville, Maryland, near Washington, D.C., and WWVH is on the island of Maui, Hawaiian Islands. Both stations broadcast continuously on the carrier frequencies and with the output powers shown in Table I. This diversity of geographical location and transmitting frequencies places the service of these stations on an essentially world-wide coverage basis.

The standard carrier frequency transmissions of WWV and WWVH are modulated with various standard audio tones and time interval signals in order to provide calibration information for a wide variety of instruments. During the first four minutes of each five-minute interval, starting on the hour, the carriers are modulated by accurate sinusoidal tone signals; 600 cycles per second is used during the first four minute period, followed by 440 cycles during the second period, etc. These two audio frequencies are alternated each five minutes throughout the hour. At the beginning of the fifth minute of each five minute interval, the tone modulation is interrupted for exactly one minute during which station identification is given in voice from WWV and in International Morse Code from WWVH. After this one minute, Eastern Standard Time is announced from WWV in voice, and Universal Time is sent in code from both stations.

During the entire transmission, a five-millisecond (.005 second) pulse signal is superimposed upon the carrier. This pulse consists of five cycles of a one-kilocycle sine wave and is heard as a sharp "click" which accurately marks the passage of each second.

This "seconds pulse" is repeated on the 89th second of each minute to accurately identify one minute intervals. Figure I illustrates the transmitting cycle of WWV which is repeated every ten minutes, starting on the hour.

New Propagation Disturbance Notices

In addition to the standard time and frequency transmissions, WWV also broadcasts radio propagation disturbance warning notices for the benefit of commercial services and amateurs whose communications depend upon conditions in the ionosphere. These so-called "alarms" are made in code during the 19th and 49th minutes of each hour. As of July 1, 1955, the system was revised to include not only the transmission of a symbol to indicate the present ionospheric conditions but also the presentation of data in a condensed tabular form.

<table>
<thead>
<tr>
<th>TABLE I</th>
</tr>
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<tbody>
<tr>
<td>FREQ. (MC.)</td>
</tr>
<tr>
<td>WWV</td>
</tr>
<tr>
<td>5</td>
</tr>
<tr>
<td>10</td>
</tr>
<tr>
<td>15</td>
</tr>
<tr>
<td>20</td>
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</tbody>
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* Reduced to 0.1 KW. for first four days following first Sunday of even months.
aspheric conditions affecting commu-
nication paths over the North At-
tantic, but also a number to indicate the condition forecast for the next
12 hours. Thus, the letters "N", "U", or "W" sent five times in code
mean that the present radio recep-
tion conditions are "normal", "unpre-
tied", or "disturbed", respectively,
while a numeral which follows each
letter indicates that the forecast for
the next 12 hours is for conditions
to continue as follows:

Forecast

1. Impossible
2. Very Poor
3. Poor
4. Fair to Poor
5. Fair
6. Fair to Good
7. Good
8. Very Good
9. Excellent

According to this system, the char-
acters "W" transmitted from WWV
would be interpreted to mean that
the present ionospheric conditions
affecting radio propagation were "dis-
turbed" but expected to improve to
"good" within the next 12 hours.

Accuracy of Transmission

The accuracy of the audio and
radio frequency and other informa-
tion transmitted from WWV and
WWVH is as great as the present
state of the engineering art will per-
mit, all frequencies transmitted
from both stations are accurate to
within 3 parts in one-hundred million.
A discussion of the means employed
to insure this precision is of general
interest and adds in gaining an ap-
preciation of the meticulous care re-
quired to maintain the national pri-
mary frequency standard.

The carrier frequencies at WWV
are derived from the average of eight
precision quartz crystal oscillators
which are operated continuously.
These oscillators all operate on 100
kilocycles per second from battery
power and are housed in resonation
vaults twenty-five feet below the
earth's surface. The temperature and
humidity of these vaults are very
carefully controlled to insure univ-
iversal frequency stability. The fre-
quencies of the oscillators are com-
pared occasionally among themselves
and are checked against the basic
frequency standard -- the period of
the second. This basic standard
"control standard" oscillator which is
chosen to control WWV drives a chain
of frequency dividers and multipliers
which convert the fundamental 100
k/c frequency to a wide variety of
standards ranging from 1 c/s to 30,
000 c/s. All audio and audio frequen-
cies broadcast are thus derived from
one oscillator and are therefore of
comparable accuracy. The 60 cycle
standard is used to drive a synchron-
ous clock motor which allows the
standard to be checked against the
Naval Observatory time. In this manner,
the time signals as broadcast from
WWV agree with Naval Observatory
time within several hundredths of a
second and have a diurnal varia-
tion which never exceeds 3 milli-
seconds.

The frequency standards at WWVH
are similar to those used at WWV and
are maintained in agreement with
WWV signals to within 2 parts in
100,000 by comparison. These
comparisons, are made during the
four minutes following each hour and
half-hour when WWVH is on the air
during 34 minute interruptions,
which occur at 1000 hours GMT.

Method of Comparison

From most points in the contin-
ental United States, the standard fre-
quency broadcasts of WWV can be
received on a relatively simple re-
ceiver. One of the superhetertodyne
type having good selectivity and auto-
nomic volume control, is to be pre-
ferred, however. These features are
usually found in communications-type
receivers. Under normal propaga-
tion conditions, such receivers are
capable of receiving WWV or WWVH
transmissions on several of the stand-
ard frequencies, thus permitting good
flexibility of measurement.

Carrier Frequency Check Points.
The simplest and most direct way of
utilizing standard frequency trans-
missions is the use of the transmit-
ted signals as check points to cali-
bate the dial of a receiver. The
variety of transmitted frequencies
usually insures that one WWV signal
falls within each tuning range of mul-
til-band receivers. The accuracy of
the receiver dial calibration may then
be checked against the standard fre-
quency carrier and any serious de-
viation corrected by adjusting either
the receiver dial mechanically, or the
receiver local oscillator trimmer. In
all cases, the receiver should be al-
lowed to reach thermal equilibrium
before comparisons are made.

Calibration of Low-Frequency R. F.
Signal Sources. Figure 2 illustrates
a method of comparing the frequency
of transmitter signals such as sig-
nal generators, grid-dip meters, and
amateur V.I.Rs with the WWV stand-
ard. The method is also applicable to fixed frequency standard oscillators which operate close to some submultiple of a WWV carrier frequency. The signal source to be calibrated is loosely coupled by radiation or capacity coupling to the WWV receiver, which is carefully tuned to an appropriate WWV signal. The frequency of the signal source is then varied until a beat note between one of its harmonics and the WWV carrier is heard in the receiver output. When the local signal is adjusted to produce zero beat, its frequency is exactly equal to that of WWV or is an exact submultiple of it. If the device being checked has an approximate calibration, the order of the harmonic is easily identified. Otherwise, the fundamental frequency of the unknown source can be found by determining the frequency differences between two adjacent harmonics on the receiver. Thus, a 1 megacycle signal will produce a beat with the 5 megacycle carrier of WWV and will also be heard at 4 and 6 megacycles, on the communication receiver, while a 3 megacycle signal will beat with the 5 megacycle carrier and will also be heard at 4.5 and 5.5 megacycles. This makes it possible to obtain many check points which will be highly accurate as long as the integral multiple is maintained at zero beat with WWV. A visual beat indicator, such as an audio output meter across the receiver output, is usually more accurate than the aural method.

Upward Extension of Standard Frequencies. The method previously described allows frequencies lower than equal to the WWV standard to be calibrated. When it is necessary to provide assured check points at frequencies considerably higher than the highest available WWV carrier, the arrangement shown in Fig. 3 is employed. In this method, a tunable oscillator of sizable frequency stability is zero beat in a receiver with the WWV standard. The harmonics of this oscillator will then appear at precisely integral multiples of the standard frequency to which it is referred. For auxiliary oscillators that have sufficient power output, these harmonics will extend quite high in frequency and may be used as "markers" throughout the VHF and lower UHF regions. The tunable auxiliary oscillator of conventional design is adjusted to operate at zero beat with WWV and supplied with unfiltered d.c. to enhance the harmonic content in order to serve for this purpose. A secondary standard of some type will usually be required to identify harmonics.

Audio Frequency Comparisons. Electronic audio equipment, as well as musical instruments capable of sustained tones, may be referred to the 440 and 600 cycle tones broadcast by WWV or WWVH by means of calibrated instruments illustrated in Fig. 4. For audio oscillators and other equipment having an electrical output signal, Fig. 4a is employed. The output of the WWV receiver (tone control and beat frequency oscillator df) is coupled to the amplifier feeding one set of plates of an oscilloscope, while the output of the device being calibrated is coupled to the amplifier feeding the other set of deflection plates. When the frequency of the local source is adjusted to equal the audio tone being transmitted by WWV, a stationary circle or ellipse will appear on the cathode ray tube. Other integral multiples or submultiples may be easily identified by means of the Lissajous figure produced. A description of the use of Lissajous Patterns may be found in a standard text.

To calibrate the musical pitch produced by non-electromagnetic sources, the equipment shown in Fig. 4a is employed. Since the source does not produce an electrical audio output, a microphone and suitable audio amplifier must be used to convert the audio output of the device into an electrical filter for comparison with that received from WWV. As in Fig. 4a, the frequency of the unknown source is identified by the oscilloscope pattern produced.

Photoelectric Cell Applications

The photoelectric cell, or "electric eye" as it is often referred to, has many applications—from use in burglar alarms and smoke detectors to facsimile, television, and even the measurement of microscopic tissue cells. It is based on a discovery by Hertz in 1887 that emission of electrons can be caused by light striking the surface.
of certain materials such as sodium and potassium.

Photoelectric devices fall into three general classes: (1) phototube or "phototubes," (2) photovoltaic cells, and (3) photovoltaic cells. Phototubes are those in which impinging light causes emission of electrons from the photosensitive surface. Most practical phototubes, such as the burglar alarm, automatic counter, door opener, and smoke detector, fall in this category. Photovoltaic cells are those in which the internal resistance varies with the amount of light striking the sensitive surface. These cells are used to operate very sensitive relays and in the measurement of infrared radiation. Photovoltaic cells are those which generate an internal emf upon exposure to light. The ordinary light-intensity meter used in photography employs a photovoltaic cell connected directly across a low resistance. This discussion is devoted to some typical applications of the various types of phototubes and photovoltaic cells mentioned above.

Phototubes

Commercial phototubes are essentially diodes contained in glass envelopes very similar to those used for thermionic vacuum tubes. The cathode is usually a large semi-cylindrical surface coated with a photosensitive material. The anode is a wire lying parallel to the cathode axis. These elements may be inclosed in an evacuated bulb, or one which is gas-filled. The gas tubes violate when the plate voltage exceeds a certain value and thus pass a larger current than do the high vacuum tubes. Gas-filled tubes are employed largely in motion picture work where their higher sensitivity reduces the amplification needed. High vacuum phototubes are used in light measurement work and in certain relay operating applications. They are less subject to damage due to application of excessive voltage or current, and their sensitivity remains more constant over a period of time.

The most common applications of phototubes involve the use of associated vacuum tube amplifiers, as in Fig. 1. The tube is coupled to the output of an amplifier by means of a large resistance, Rg. Since the current flow through the cell is of the order of a few microamperes, this resistance should be very high. By proper amplifying circuits, the current in the final output stage of the amplifier may be sufficient to operate a relay or a loudspeaker as in the sound picture industry. See Fig. 2.

Another valuable application of the phototube is the control of lighting. The tube is used with an amplifier and relay to turn the lighting system on when daylight decreases and off when natural light is again adequate. Fig. 3 illustrates a circuit in which the relay is energized by an increase in light. As long as the illumination on the phototube is below a certain value, the 3051 grid potential is below cutoff, and prevents
conduction. When illumination rises, grid voltage is made less negative and the tube conducts, closing the relay. The function of R4 is to keep the current through the 2051 within the tube maximum rating. Note that this circuit works directly on a.c. line voltage, requiring no d.c. supply.

Photocell controls

The simplest use of the phototube and relay is that of counting. A beam of light is directed across a conveyor belt into a phototube which operates a counter. When the beam of light is interrupted by one of the objects to be counted, the change in tube current operating the counter.

An interesting circuit of this type is the one-ray counter illustrated in Fig. 4. This arrangement records objects passing in one direction, but not in the other.

Suppose an object is passing downward in Fig. 4 so that it obscures phototube A and then B. When the light to tube A is interrupted, plate current flows in tube X, opening the contacts of relay X. As the object continues downward, both tubes are obscured and relay Y closes. But since the contacts of X relay are open, no current flows through the Y relay and the counter is inoperative. Now suppose that the object passes from B to A. Relay Y is operated when amplifier tube B starts to conduct. Then, when the object clears both phototubes, the current through the amplifier tube associated with phototube A passes mainly through the contacts of relay X and Y to operate the Y relay and the counter. Relay X does not operate and its contacts remain closed. Thus, the counter is actuated only by objects passing in the direction from B to A.

Industrial safety controls

The applications of photocell devices to safety devices are very numerous. Some of the more familiar safety controls are the smoke detectors, traffic control, gate protective devices, bar-type switches, and fire alarms. Another important protective circuit of this type is the flame-failure detector shown in Fig. 5. This device, sensitive to flashes of light, is used in protecting the oil vapor in the tank from explosion. The flash detector shown in Fig. 6 is used to protect the oil furnace, using a dual time as its principal element. When light from the flame is present, photocell flows and the first time section is blocked. The second section normally conducts current but to close the relay which opens the steam oil valve and allows the flame to burn. Should a flame failure occur, the photocell no longer provides blocking voltage to the first section, which then conducts and applies a blocking voltage to the grid of the second triode section. The blocking of current in the second time opens the relay and closes the oil valve with the simultaneous ringing of an alarm bell.

An even more common kind of industrial use of a photocell is the "light curtain" type of protective device used to safeguard the operators of heavy machines. In this application of phototube devices, a light curtain is formed about the area of danger by a series of beam projectors and mirrors, the beam falling ultimately on a set of phototubes. If the operate...
Printed Electronic Circuits

The reproduction of electrical circuits on insulating surfaces by various printing techniques has become a small, lightweight, economical electronic option to elaborate assemblies that are costly to manufacture and can be used on miniature equipment. The use of printed circuits is no longer confined to a few military devices and hearing aids, but may now be encountered in a large number of everyday equipment. These include speech amplifiers, portable television receivers, two-way radios, television receiver front-panels, FM receivers, and many others. For this reason, a working knowledge of the design, production, and maintenance of such circuits will be a valuable asset to any worker in the electronics field.

This discussion is concerned with the general types of printed circuits, the relative advantages of each, and methods of effecting servicing repairs. The use of printed circuitry has been revolutionary not only because it permits the fabrication of extremely small and inexpensive electronic components, but also because it reduces the production of such components to a simple, rapid operation which is almost completely devoid of the possibility of human error. By this method, a relatively unskilled operator can reproduce literally hundreds of complex units in the time formerly required to make one unit by old-fashioned "wire-wrap" soldering techniques. In addition to electrical conductors, critical visual components such as resistors, capacitors, and inductors can be "printed" into the circuit in the same operation and held to close, reproducible tolerances. Figure 1 shows a typical printed circuit and its schematic diagram.

![TYPICAL PRINTED CIRCUIT AND DIAGRAM](fig1.png)

90
Printed circuits are classified ac-
cording to the method of pro-
ducing them. These are, at present, six general types. The processes are: painting, spraying, vacuum evapor-
ing, chemical etching, metal-stamping, and powdered metal dipp-
ing. Each of these general categories will now be discussed in some detail.

**Printing Techniques**

Probably the most widely used pro-
cess for producing printed circuits is the painting technique. In this meth-
od, the conductors and other compo-
nents of the circuit being fabri-
cated are painted on the insulating surface which acts as the base for the circuit.

The paint may be applied by hand with a brush, although in production operations the silk-screen stenciling process is more frequently used. Thin ceramic or plastic sheets may be em-
ployed for the base, or a metallic sur-
face covered with an insulating lac-
quer may be used. In special in-
stances, the glass envelope of a vac-
uum tube has been utilized as a base for its associated printed circuit. See Fig. 2.

The paint used for electrical con-
ductors consists of a powdered metal sus-
cept to copper or silver in suspension in a liquid binder. This conducting paint is applied to the surface of the insulating material through metal stencils. Other paint, made up of a resistive material, such as carbon, may be applied in specific amounts to form resistors. Capacitors may be made by painting the plates on opposite sides of the base plate, if the required capacitance is small.

Otherwise, small capacitors are con-
ected to the printed circuit as in Fig. 3. It is interesting to note that these ca-
pacitors are manufactured by processes which are essentially printed circuit tech-
niques. Inductances are produced by painting serpentine of conducting paint on the surface of the ceramic or other base material. "Crosswires" in the spray-
ing are made by painting one conductor director backward, and another with a layer of insulating material such as lacquer be-
tween the two, or by depositing a metallic deposit on the surface of the base plate and then, after painting the metal strips, deposit the two metal strips on the other side of the plate for a short distance. The metal strips are then made a part of the printed circuit by etching the metal strip into the insulator, as illustrated in Fig. 4.

When all printed components have been painted in place, the entire as-
sembly is fired at an elevated tem-
perature to fuse the metal particles together and bond the circuit to the base plate. Temperatures ranging from room temperature for plastic bases to as high as 800 degrees C. for ceramics are used.

Vacuum tubes, external leads, and other components not printed are

soldered to eyeslets in the base plate, as in Figs. 1 and 3. To take maximum advantage of the space-saving prop-
erties of printed circuits, tubes of the subminiature type are usually em-
ployed.

The painting technique has the adv-
antage of requiring a minimum of auxiliary equipment and so has been the most popular type for experimenta-
tion and design work with printed circuits. It is also the best method to use in making repairs on printed circuits, as will be discussed later.

The spraying method of applica-
ing printed circuits differs from the painting technique in that the con-
ductors are sprayed onto the surface of the base. Both metallic metals and metallic conducting paints may be ap-
plied in this manner. In some pro-
cesses, stencils are used to define the circuit conductors. In others, grooves are machined or pulled in the base material where a conductor or other circuit component is desired. Grooves may also be formed by sand-blasting through a stencil. Metal is then spray-

ed vaporized from a heated element, or other source of metal vapor, over the printed circuit plate placed on a substrate for vacuum processing, it is unnecessary to further heat treat or to deposit the metal. Only thin films are usually formed. Stencils are manufactured by the use of fine wire screen printing techniques. Conductivity is required, conductors may be built up by electropainting.

In the application of the various methods of making printed circuits, the tech-
niques employed are similar to those used in silvering mirrors. A silvering solution, consisting of ammonia and silver nitrate mixed with a reducing agent, is spread over the surface, dried, and then baked to prevent surface to be coated. The en-
velope of the solution are controlled by the thickness of the material obtained. This material may be built up by repeated coatings or by pott-
ing. The chemical processes have not been applied as extensively as those discussed above.

The metal stamping technique has been used principally to print loop antennas on the back sides of radio receivers. However, other types of circuit wiring have been produced by this method. A die, bearing the cut-

line of the desired circuit, is used to press a tin metal foil into the sur-
face of a plastic or another insulator. In
the same operation the sharp edges of the die cut the metal sheet to the desired shape. The metal sheet may be backed by an adhesive to insure a good bond. Circuits made in this manner have good conductivity.

The most general type of printed circuit is produced by a process known as "stamping". In this method, a powdered metal is dusted onto the insulating base plate and fired in place. The circuit outline is defined either by coating the entire insulator with a sticky substance and applying the metal powder through a stencil, or by applying the bonding solution through the stencil and dusting onto the powder so that it is held in place by the adhesive until fired.

Servicing Printed Circuits

As was mentioned above, the most convenient method of making repairs and replacements in printed circuits is the brush-applied painting technique. Kinds of such paints, including both conductor and resistor mixtures, are commercially available. Most of these paints require no heat for drying, so they may be used for repairing circuits having parts which cannot be subjected to high temperatures. This is an important precaution when working with circuits printed on certain types of plastic. Although subminiature tube sockets are sometimes used with printed circuits, tubes are frequently connected directly to metal eylets or the base plate, as in Fig. 1. When replacing tubes connected in this manner, care must be exercised to avoid the use of excessive heat during soldering operations. Soldered connections may also be made directly to printed conductors if the base material will stand the heat involved. A solder container containing a small percentage of silver should be used for best results. Where soldering is unavoidable, connections to tubular sockets, and other wires should be made with metallic paint.

Printed resistors which have become defective may be repaired or replaced by the painting technique. Depotive resistors are located in the usual manner with its ohmmeter. If it becomes necessary to "disconnect" a printed resistor from the circuit for a resistance check, this may be accomplished by scratching through the printed conductor lead with a sharp instrument. If defective, the resistor may be repaired with resistive paint. It will usually be found to be open or high in value. In such cases, additional resistive paint should be applied over the old resistor to reduce its resistance to the proper value. Some commercial printed circuits have a protective layer of lacquer over the conductors and particularly over resistors to prevent moisture absorption. This coating must be completely removed before replacing resistors. It attempts to repair defective resistors are unsuccessful, the old coating should be removed completely and a new resistor painted in its place. The proper dimensions may be determined by trial and error, keeping in mind that the resistance is directly proportional to the length, and inversely proportional to width and thickness. The resistance material must make good contact with the printed conductors at the ends. Breaks introduced in the conductors to isolate resistors may be repaired with a bridge of conducting paint.

The Transistor - An Amplifying Crystal

Among recent technical developments, the "transistor" or semiconductor triode, will probably have the greatest effect upon every field of electronics. Although still in its infancy and by no means comparable to that represented by the De Forest "audion" in the development of the vacuum tube, the implications of this tiny, heatless, vacuumless capsule are so tremendous as to warrant the attention of every worker in the radio field.

Just as the unidirectional conduction of current in the thermionic diode (Böhm effect) was known for over 24 years before the addition of a third controlling element in the form of a grid, the rectifying properties of such semiconductors as galena, iron pyrites, silicon and germanium have been used in radio applications for many years. During this time very little thought was given to the possibility of electronically controlling such rectification. However, due to impetus gained through the widespread use of crystals as microwave mixers and detectors during the last war, recent research in the field of semiconductors has resulted in the discovery of the crystal triode, or Transistor.

Physically, the transistor consists of a small crystal diode, the size of a pin head, on which is mounted the transistor housing. As was used in great numbers during the last war, the emitting crystal has been modified only by the addition of a second "collector" crystal. This makes contact with the germanium semiconductor at a point very close to the point of contact of the first cathode; the second crystal acts as a shield to protect it. It is the addition of this element - the device which permits control of the current flowing to the first and enables it to perform many of the functions of a vacuum tube triode, although in a different way, as will be shown.

The construction of the transistor, as pioneered at the Bell Telephone Laboratories is shown in Fig. 1. This mounting closely resembles the "coaxial cartridge" type commonly used in rectifier crystals, except that two
center conductors are used. A small rectangular, ‘waterproof’ semiconductor (germanium crystal) is soldered to a brass disc, which in turn is fitted into one end of a small metal tube approximately 3/16 inch in diameter and 3/4 inch long. Thus the device is completely in a half-watt resistor. The two call-whiskers are of tungsten wire approxi-
mately .002 inch in diameter wedged to the ends of rigid wire terminals which lie in turn as spaced by an insu-
lation board. This supports them with-
in the metal tube so that both call-
whiskers are brought in contact with the highly polished surface of the germanium. The separation be-
tween the points of contact is main-
tained at between .002 and .005 of an inch. The structure is ruggened against mechanical shock and vibra-
tion by impregnating the capsule with a low-loss compound which also renders it moisture proof. It is prob-
able, that, as the development of this mechanically simple device progresses, new applications will dictate radi-
cal departures from this preliminary design.

Experimental applications to which the transistor has been adapted in-
clude: radio-frequency amplification,
stages, audio and i.f. oscillators, in-
termediate frequency amplifiers, audio amplifier stages and other types of circuits constantly employing
vacuum tubes. As yet, the upper frequency limit of the transistor is about 10 megacycles. The reason there appears to be no fundamental reason why this range of usefulness cannot be extended into at least the VHF range, to 150 megacycles and f.m. applications. However, transistors are operating in con-
ductor devices as in negative grid voltage amplifiers. The peak of operation beyond the VHF range. Although transistors are not as efficient as detectors and converters of microwave energy at frequencies exceeding 25,000 megacycles/second, the type of high-inverse-voltage ger-
manium semiconductor used in the present transistor is not an efficient rectifier at frequencies much above 60 megacycles/sec.

Another limitation to the use of the transistor in some circuits, appli-
cations is the excessive noise voltages generated within it at present. This noise is about 70 db above the theo-
retical thermal or Johnson Noise, which is defined as the noise that would be generated in an equivalent resistance due to the random motion of thermally agitated electrons with-
in it. This objectionable noise char-
acteristic is most pronounced at the audio frequencies and appears to de-
crease somewhat with frequency.

**FIG. 2**

Future laboratory work will undoub-
tedly result in methods of reducing
such noise to more usable levels. If noise figures comparable to those en-
countered in the modern vacuum tube are achieved, the transistor will re-
place tubes in many low-power ap-
plications, since it consumes no heat or filament power, requires no warm-
up time, is virtually heatless in oper-
ation and is more compact and rugged than even the smallest vacuum tubes. Because of these advantages, transistors will find use in circuits where economy of power consump-
tion and extreme compactness is de-
sirable. Such circuits may include ultra-portable broadcast receivers using printed wiring and components, electronic computing devices which may require as many as 18,000 vac-
uum tubes (as in the ENIAC), elec-
tronic musical instruments which also use hundreds of vacuum tubes oper-
at ing at low audio levels and other
more conventional applications.

The more fact that the transistor does not require a vacuum is an important ad-
vantage, since the production and maintenance of a satisfactory high
voltage vacuum is an item of major expense in the manufacture of electron tubes. In addition, the operating efficiency of the transistor exceeds that of vac-
uum tubes in many applications, es-
pecially when it is considered that no heater power input is required. Op-
erating efficiencies as high as 25% have been observed at output levels as high as 20 milliwatts.

The basic electrical circuit of the transistor is shown in Fig. 2. The in-
put electrode, which is called the ‘emitter’, is maintained at a small positive potential with respect to the
 germanium block. The impedance to current flow in this direction is low (100-400 ohms) due to the rectify-
ing action of the germanium-tungsten contact. Therefore, this small posi-
tive “bias” voltage (.1 to .5 volt) causes an appreciable “forward” cur-
rent to flow in the emitter circuit. Also, because of the low impedance to forward currents, a small incre-
ment in emitter voltage causes by an impress signal will result in a
large increase in electron current flowing from the semiconductor to the collector. The static voltage-
current characteristic of the emitter circuit, when considered alone, is similar to that of the typical ger-
manium point-contact rectifier, which is shown in Fig. 3. On the other hand the ‘collector’, or output circuit of the transistor, is biased negatively with respect to the germanium. At this polarity, the impedance to cur-
rent flow is relatively high (exceed-
ing 10,000 ohms), so that over 30 volts may be applied to the collector before appreciable “back” current is from 10 to 20 milliamperes.

The dotted portion of Fig. 3 represents the static (no load) characteristic of the collector circuit in the absence of the emitter. The close proximity of the two characteristic curves is re-
spective operating voltages modifies these characteristics considerably, how-
ever. It is the ability of the
transistor to transfer an emitter volt-
age change to the collector circuit in a manner which gives the device its name, which is the basis of the transistor as an amplifier. This property results in effective power gain of 100 times or 20 db being possible.

To understand the mechanism wherein the emitter circuit power varia-
tions induce relatively large power variations in the collector circuit and hence, enables the transistor to ampli-
ify it is necessary to examine the fundamental properties of semicon-
ductors. Although the theory of the transistor is still not completely un-
derstood and may be subject to fre-
quent revision, a working concept of its operation can be gained from a simplified analysis and by drawing analogies to vacuum tube circuits.

It will be recalled that the resist-
rance to current of a number of free electrons available in the molecular structure for the conduction of current. In some
materials the electrons in the outer "shells" or orbits of the atoms are so loosely bound to the nuclei that they can be easily removed and may move about between atoms under the influence of thermal agitation. If an external voltage is applied, such as the conduction of a battery across such a metal, the random motions of these free electrons within it are coordinated into a stream of electric current flow. The metals which may have as many as one to two free electrons per atom, are examples of such insulators which are called conductors of electricity. Other substances, on the other hand, may have virtually no free electrons available for current conduction and are classed as insulators. The semiconductor bulk is intermediate between metals and insulators in the scale of conductivity, having perhaps only one free electron for every millions of atoms.

The extent and type of current conduction possible within them is dependent upon the very minute quantities of impurities present. Thus, although perfectly pure germanium, if obtainable, would be a very poor conductor of normal temperatures, the addition of one part ten million to fifteen parts per million of impurities will impart conductivity to it.

The word "impurity" is used here since current in a semiconductor can be carried in two different ways: by the presence of free electrons within it, or by the absence of some of the "bound" ones. The latter type of current is called hole current. As the material physiologically has an exactly conduction by holes, it is a semiconductor. There are three types of impurities in the otherwise filled lines of valence electrons which can create a type of current flow. One is the type of conduction which can be changed by heat-treatment.

In the transistor, both n-type and p-type current conduction are present. The emitter is biased positively to attract free electrons which surmount the rectifying barrier layer at the interface of the semiconductor-base contact and flow in the input circuit. The collector, biased negatively, attracts a small number of "holes" in the absence of emitter current. The useful property of the transistor arises from the fact that a flow of electron current to the emitter collector-up embraces the formation of many more electron vacancies or "holes" at the surface of the semiconductor which are gathered by the nearby collector cathode. The result is an increase of output current. A change in the emitter circuit current "modulates" the number of "holes" available to the collector and thus causes a similar change in collector current. Although the changes in the input and output circuits are of the same amplification, amplification results since the collector circuit impedance is approximately 100 times that of the emitter circuit. Fig.

[FIG. 4]

VACUUM TUBE CIRCUIT WHICH HAVES CHARACTERISTICS SIMILAR TO AN ACTION OF A TRANSISTOR IS SHOWN IN FIG. 5. In this circuit the transistor is replaced by a dual-diode vacuum tube which has a small quantity of residual gas. The two small anodes are spaced close together and are biased by voltages of similar polarity to those applied to the two contacts of the transistor. The anode which is at a small positive potential with respect to the cathode has characteristics which are comparable to the emitter electrode; it attracts electrons from the cathode and thus the circuit presents low impedance to "forward" current flow. The diode plate which is biased negatively is analogous to the collector cathode of the transistor in that the circuit impedance is very high and little current flows in it normally. However, the flow of current in the "emitter" may be made to increase the current flowing in the collector circuit. Electrons being accelerated toward the emitter anode may collide with residual gas molecules, remove electrons from them, and thereby form positive ions which are drawn to the negative collector anode. These ions are roughly analogous to the "holes" in the transistor case since they behave like positive particles and move more slowly than electrons. The number of ions formed depends upon the number of electrons flowing to the emission electrode. Therefore, when the input circuit controls the flow of current in the output circuit, just as in the transistor and conventional vacuum tube amplifiers.
The advent of the transistor opens fertile fields for future development. Much more can be done in the adaptation of conventional circuitry for use with this simple device. Problems frequently arise since the electrode voltage polarities and circuit impedances differ so greatly from those encountered in vacuum tubes.

The cascading of amplifier stages is complicated by the fact that the input and output impedances do not lend themselves well to the usual schemes of interstage coupling. The possibility of push-pull and parallelized units for increased power output is still to be fully exploited.

Because of its simple structure and the small amount of test equipment required, the transistor provides a valuable subject for individual experimentation. Transistors have been fabricated for study purposes from germanium removed from standard 1N34 diode crystal rectifiers suitably mounted with two adjustable fine wire cat-whiskers.

### Junction Transistor Circuits

The former tightness of the transistor situation has been eased favorably for the non-military experimenter and manufacturer by the present excellent availability of the new junction-type transistors. The lower price of these components, compared with the almost prohibitive price of the earlier point-contact transistors, should stimulate private development of transistor circuits. It is expected that prices will drop further in proportion to the number of circuit applications which can be developed to utilize the wide spread in coefficients resulting in transistor manufacture.

The circuits included in this discussion have been made to work satisfactorily and can be duplicated. It should be borne in mind, however, that these circuits satisfied one set of typical conditions and do not necessarily represent the best or only way of applying the transistor for the purpose intended. Considerable flexibility in individual design is possible. In addition, some readjustment of constants may be required when transistors of various manufacturers are used. The circuits described are intended especially for junction-type transistors and some of them often will not operate equally well with point-contact transistors. In presenting this material, we feel that it will be invaluable in guiding the newcomer to transistor circuitry and will be of provocative importance as well.

### Features of the Junction Transistor

Several characteristics of the junction transistor distinguish it from the point-contact type. One of the most important of these is the increased ruggedness of the junction type. In the junction transistor, the three conduction layers (E, N, and P in the case of the C722) are parts of the same germanium wafer. There accordingly are no whiskers or sand with sections which might be displaced accidentally.

A dramatic property of the junction transistor is its high efficiency.

### Absolute Maximum Ratings

- Collector Voltage (Vc) = 20 volts
- Collector Current (Ic) = 3 ma.
- Collector Dissipation 30 W. at 30° C.
- Emitter Current (Ie) = 5 ma.
- Ambient Temperature 50° C.

### Typical Collector Characteristics

![Typical Collector Characteristics](figure_1)

### C722 Operating Data

**Collector Voltage (Vc):**
- 1.5 volt

**Collector Current (Ic):**
- 0.5 ma.

**Base Current:**
- 20 µa.

**Current Amplification Factor (β):**
- 12

**Power Gain:**
- 1000 (30 db) Source 1000 e.m.s.; Load 20,000 e.m.s.

**Noise Factor:**
- 22 db at 1000 cycles

*This rating applies only to the grounded-emitter circuit. The current amplification factor α for the grounded-base connection is, of course, less than 1 for the junction transistor.*

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95
and its ability to operate at very low values of applied d.c. voltage. A class "A" amplifier using a junction type, for example, will operate close to the theoretical 50% efficiency point, as compared with a vacuum tube amplifier giving 25 to 30 per cent. Practical amplifiers and oscillators can be operated from a single 1½-volt cell with current drain so low that in some arrangements the cell will last almost indefinitely. Audio oscillators can be made to operate at such low d.c. levels that, in demonstration, the "power supply" current has been furnished by a self-generating phonet, thermocouple, or makeshift wet cell made from two coins separated by a piece of paper moistened with saliva.

The temperature sensitivity of the junction transistor makes the latter somewhat poorer than the point-contact type, but the junction type is not as noisy. The maximum ambient temperature allowed for the CK722 is 50°C. The 1000-cycle noise factor is 22 db. (Compare the noise factor of 65 db. which is given for the CK116 point-contact transistor). Frequency response of the junction transistor appears to be lower than that of the point-contact type and is limited by such factors as the increased capacitance of the junction layers and the differences in mobility of the carriers. Our tests indicate that the CK722 is suited particularly to audio and low-frequency r. f. applications, of which there are many in each category. As a radio-frequency oscillator, this unit has given good performance in our circuits at high as the upper limit of the standard broadcast band, but beyond that point its operation has not been encouraging.

Figure 1 shows a family of collector current-voltage curves for the CK722. These curves are plotted for eight values of constant base current (0, 5, 10, 20, 30, 50, 100, 200). Note that these curves have the general appearance of periodic vacuum-tube curves. The collector voltage (Vc) values are negative. The corresponding collector currents (Ic) also are designated as negative.

The Table in Figure 2 lists important operating data for the CK722. One listing is apt to confuse the reader who has read some prior contact with transistor literature. This is the current amplification factor, always mentioned as less than unity for junction transistors, which is given here as 12. The reason for this higher figure is in the factor given in Fig-
ure 2 is not alpha (which is less than 1) but beta which applies only to the grounded-emitter (low-input) operation shown. Ref (b) is related to alpha (a) approximately as follows: \( b = 1/(1-a) \).

Junction Triode Circuits

Figures 3 to 9 show several selected amplifier and oscillator circuits. These preliminary circuits serve as building blocks for more complex equipment. Note that each of these arrangements uses the low d.c. voltages at which the junction transistor is capable of operating.

Single Amplifier Stages. Figure 3 is a resistance-coupled, grounded-base audio amplifier circuit. The grounded-base arrangement is the progenitor of all transistor circuits.

The grounded-base circuit has an input impedance of approximately 1000 ohms and an output impedance of 5000 to 10,000 ohms, depending upon individual transistor collector characteristics. Higher operating impedances are possible in the output with higher \( R_b \) values, but with somewhat reduced gain. Operating into a high-impedance load (100,000 ohms or higher), this stage, as shown, has a voltage gain of 40, although the gain may vary between 26 and 44 with individual transistors. At lower load resistance values, the gain drops proportionately.

With 1-microfarad input and output capacitors \((C_1\text{ and }C_2)\), the frequency response is such that the gain of 100 cycles is 25% of the 1000-cycle value, and at 20,000 cycles is 95% of the 1000-cycle value. With 10-microfarad capacitors, the 20-cycle gain is 70% of the 1000-cycle value, and the 100-cycle gain is 90% of the 1000-cycle value. Miniature, low-voltage electrolytic coupling capacitors may be used for the high values.

Because the grounded-base amplifier requires two batteries, there is some objection to its use. Current drain of the emitter battery is 150 microamperes, and of the collector battery 100 us. The grounded-base amplifier offers the maximum power gain possible with a given transistor.

Figure 4 shows a grounded-emitter amplifier. An important advantage of this circuit is its ability to operate with a single battery at a drain of 10 or 50 microamperes, depending upon the individual transistor employed. Input impedance is of the order of 1000 ohms; output impedance 20,000 to 40,000 ohms. Higher output impedance values are possible with higher values of \( R_b \) but with reduced gain.

With the constants given in Figure 4, voltage gain of this stage is 40 to 50 when \( B \) is 1 1/2 volts, and 80 to 100 when \( B \) is 3 volts. These gains are obtained only when the stage is worked into a high load impedance (100,000 ohms or higher).

Frequency response is the same as that quoted for the grounded-base amplifier in the foregoing paragraph.

Figure 5 shows a grounded-collector amplifier. This circuit has high input impedance (of the order of 50,000 ohms) and low output impedance, 1000 ohms. It is thus equivalent to the cathode-follower vacuum-tube amplifier. Like the cathode follower, the grounded-collector circuit provides no voltage gain ("gain" of the stage shown in Figure 5 is 0.2 to 0.3). It does afford power gain, however, of the order of 15. The frequency response of this stage is the same as that stated earlier for the grounded-base circuit.

A slight disadvantage of the grounded-collector type of circuit is its requirement of two batteries \((B_1\text{ and }B_2)\). But its relatively high input impedance makes it very well to use as the input stage of a transistor amplifier whenever the loss in voltage gain is of no consequence.

We did not discover that bypassing, either of the power supplies in any of the circuits shown afforded improvement in performance at any frequency between 20 cycles and 20 kc.

Cascaded Amplifiers

The circuits given in Figures 3 to 5 are fundamental building blocks. Like tube circuits, transistor amplifiers may be cascaded for increased voltage gain and power gain. The difference with transistors, how-
ever, rests in the fact that in grounded-base and grounded-emitter stages, the output impedance is higher than the input impedance. This requires an impedance stepdown between stages. While resistance coupling may be employed as well as transformer coupling between stages, the greater power gain will be obtained with interstage transformer coupling, since the latter has the lower stepdown ratio and offers better power transfer. In resistance coupling, at least one additional transistor stage usually is necessary to provide the same overall power gain afforded by transformer coupling.

Figure 6 shows one method of resistance-coupling three junction transistor stages. Overall power gain is approximately 60 db. Collector resistors R5, R6, and R7 each is 20,000 ohms. Base resistors R1, R2, and R3 each is 150,000 ohms. For best results, each of these resistors should be adjusted carefully for best gain and lowest noise output with the individual transistors used. Capacitors C1, C2, C3, and C4 each is 10 microfarads. This amplifier will deliver approximately 2½ milliwatts output to a high-impedance load. A 1000- or 2000-ohm headphone may be connected in place of R6 and C4 to obtain approximately the same output in such applications as hearing aids, pocket radios, receivers, etc.

Figure 7 shows a transformer-coupled 2-stage transistor amplifier. This unit has an overall power gain of approximately 35 db. The interstage transformers have primary impedances of 20,000 ohms each, and secondary impedances of 1000-ohms each. For experimental setups, good results can be obtained with carbon-microphone transformers connected backward. The output transformer has a 20,000-ohm primary. Its secondary may have the proper value required to match a small loudspeaker, headphone, line, or other device. If desired, a 1000- or 2000-ohm headphone may be connected in place of the primary of the output transformer, T5. Suitable subminiature transformers for use in the transistor amplifier intended for hearing aids, pocket receivers, are available at most parts distributors. A high-impedance crystal microphone may be coupled into the first transformer by using a 200,000- to 1000-ohm input transformer at T5.

In Figure 7, capacitors C5 and C6 each is 10 microfarads. Resistors R1 and R2 each is 150,000 ohms. Using the fundamental building blocks, a number of combinations of cascaded amplifier stages is possible to suit individual requirements. For example; grounded-base, grounded-emitter, grounded-collector, resistance-coupled, and transformer-coupled stages may be combined, as needed.

Transistor Oscillator Circuits

The CK722 junction transistor appears to oscillate most readily in an
inductive-feedback type of circuit. Figure 8 shows two audio-frequency oscillators employing this principle. Figure 9 is a radio frequency oscillator employing inductive "feedback". 

Audio transformers are used in Figure 8 (A) and (B). In each instance, the high-impedance winding is connected to the collector. A satisfactory transformer is the type used to couple a single triode plate to 500- or 600-ohm line. Satisfactory results may be obtained also with a carbon-microphone transformer. The transformer must be phased properly for oscillation. If oscillation is not obtained immediately upon application of battery voltage, reverse the connections of either the primary or secondary. With a microphone transformer at Y in each circuit, a 750-cycle signal was generated. The "natural" frequency will depend upon the inductance of the windings and their distributed capacitance, and may be lowered by means of capacitors connected at C. 

Figure 8 (A) shows a grounded-base oscillator; Figure 8 (B) a grounded-emitter oscillator circuit. The first circuit requires two batteries but is somewhat less temperature-sensitive than the second. Air-wound coils are used in the radio-frequency oscillator. Figure 9. The top frequency at which this circuit has been operated with the CR722 is 150 kc. No frequency data are published on this oscillator.

Tight coupling is employed between coils L\textsubscript{1} and L\textsubscript{2}, the former being wound on top of the latter. The output coupling coil, L\textsubscript{3}, is wound on the same form close to L\textsubscript{2}. By making these coil sets plug-in, frequency bands between 50 and 1500 kc may be covered.

A good broadcast-band oscillator may be made with L\textsubscript{4} a 540-1750 kr antenna coil. L\textsubscript{5} is the slip-on primary normally supplied with the antenna coil. L\textsubscript{6} consists of 25 turns of No. 30 enamelled wire closewound on top of the manufactured coil. L\textsubscript{6} coil L\textsubscript{3} is insulated from L\textsubscript{6} with Scotch tape. C\textsubscript{3} is a 350-uf. tuning capacitor.

An interesting regenerative broadcast receiver having good sensitivity can be made by connecting antenna and ground to the two terminals of L\textsubscript{5} and a pair of 2000-ohm (or higher, magnetic) headphones in series with the collector and L\textsubscript{6}. Regeneration can be controlled by means of a 1-megohm potentiometer substituted for the 220,000-ohm fixed resistor shown in Figure 9. A transistor radio amplifier may be added by substituting the amplifier input transformer for the 1-megohm potentiometer. Near the vicinity of strong local stations, an outside antenna and ground are not required, an ac-dc antenna tank, connected to one terminal of L\textsubscript{5} being sufficient. The other terminal of L\textsubscript{5} then would be connected to positive terminal of the battery, as shown in Figure 9.

Class-B Transistor Amplifier Data

The transistor is an efficient device, vice by nature. Even when we fail to take into account the absence of filament power, the transistor is found to offer higher operating efficiency than the vacuum tube. This is especially true of the junction transistors.

In transistor amplifiers, as in tube amplifiers, higher efficiencies are made possible by class-B operation. There are two main reasons for the present growing interest in class-B transistor operation. First, class-B gives maximum output power per dollar of initial cost—an important consideration when utilizing present, high-priced power transistors. Second, the rather low power output of conventional transistors may be boosted several times by utilizing class-B.

Class-B Tube vs Transistor

It is a conventional class-B tube amplifier, the control grids are biased for plate current cutoff, or very nearly so, under zero-signal conditions. The plate current then is driven to a certain peak value at maximum signal. Large economies are effected by the low, resting, zero-signal d-c plate power input.

There are two ways of adjusting a transistor amplifier for class-B operation. Under zero-signal conditions, the output electrode (usually the collector) may be operated either at low direct current and high voltage (comparable to static operation of the class-B tube amplifier) or at high direct current and low voltage. In the first instance, the input e, c signal will drive the output. In the second case, it will drive the current downward.

The family of common-emitter collector curves in Figure 1 will serve to illustrate this point. When the circuit is adjusted for zero-signal operation at point a, the resting point is X. The positive half-cycle of the input signal then will drive the collector...
current from $i_2$ down to $i_c$, and the operating point to $Y$ which corresponds to $v_{IC}$. When, instead, $Y$ is chosen as the no-signal operating point corresponding to $v_{IC}$, the negative half-cycle of the input signal will drive the collector current up from $i_c$ to $i_b$ and the operating point to $X$ which corresponds to $v_{IC}$.

While both operating points ($v_{IC}$ and $v_{E}$) can represent points of low “resting” collector dissipation, the greater overall operating economy is affected by the high-voltage, low-current condition ($V$, $v_{E}$). This is because the establishment of the opposite condition, a low collector voltage at high current from a constant-voltage collector d. c. supply would necessitate use of a dropping resistor with attendant I.R. loss. Large initial collector current also reduces the transistor current amplification factor, $\alpha$.

As in the class-B tube amplifier, dynamic output-electrode current in the transistor class-B amplifier becomes a series of quasi-half-sinusoids. Thus, in the common-emitter circuit with static characteristics such as displayed in Figure 1, the half-sinusoids of collector current would extend from the “zero” value, $i_c$, to the peak value, $i_B$, and back.

Figure 2 shows a comparable family of curves for the common-base transistor amplifier configuration.

The currents and voltages $i_1$, $i_c$, $i_v$, $v_C$, and $v_V$ are d. c. or peak signal voltage, or downward from point X by negative half-cycles of emitter signal voltage. This is the opposite of conditions with the common-emitter configuration.

Circuit Configurations

Transistors must be used in push-pull pairs in class-B amplifiers, the same as in comparable tube amplifiers. However, the distortion reducing properties of the asymmetrical arrangement are somewhat less in the transistor circuit. Either of the three well-known transistor circuit configurations may be employed: common-base, common-emitter, or common-collector.

D. C. output-circuit efficiency runs close to 80 percent for each circuit. Power gain is highest with the common-emitter, less by a factor of 10 with the common-base, and quite low (of the order of 10) with the common-collector. However, the common-base configuration affords the highest power output and overall power gain, for a given distortion level. The output vs distortion ratio is due to its more favorable $v_{IC}$ characteristic.

Figure 3 shows typical circuit arrangements for the three configurations. While batteries are shown for simplicity, the bias voltages may be obtained likewise from a. c. — operated power supplies.

The currents and voltages $i_1$, $i_c$, $i_v$, $v_C$, and $v_V$ are d. c. or peak signal voltage, or downward from point X by negative half-cycles of emitter signal voltage. This is the opposite of conditions with the common-emitter configuration.
values obtained from the static characteristic curves of the transistor employed. For example, in the common-emitter circuit (Figure 3b), \( i_1 \) corresponds to collector current \( i_c \) at point X in Figure 1, and \( i_1 \), to the constant base current \( i_1 \) in Figure 1. The base voltage \( (v_b) \) and emitter voltage \( (v_e) \) must be determined experimentally for the particular transistor employed, since most transistor manufacturers do not now supply input characteristic curves. This is done by finding the base voltage in a common-emitter circuit which will produce base current value \( i_b \). (Figure 2.) The emitter voltage which will produce emitter current \( i_e \). (Figure 1.) In each case, it is assumed that collector current is held constant at value \( i_c \). (Figures 1 and 2) during the measurement.

Design and Operating Data

After selecting the type of junction transistor to be used and obtaining a pair matched for \( v_i \), characteristics and \( \alpha_1 \), the first requirement is choice of collector supply voltage, \( v_c \). This voltage must not exceed one-half the maximum permissible peak inverse voltage specified by the transistor manufacturer.

In most applications, the common-emitter circuit will be employed, because of its superior power gain. The maximum collector supply voltage then corresponds to \( v_i \) of Figure 1, and this value is to be \( v_i \), the maximum peak inverse. The point \( X \) must be selected such that the product \( v_i \), does not exceed the maximum permissible collector d. c. power dissipation specified by the transistor manufacturer. The primary winding of the output coupling transformer, \( T_0 \), is assumed to have low d. c. resistance, \( r_o \), in order to minimize voltage drop between \( v_i \) and the collector.

The following approximate common-emitter class-B design equations have been adapted from those given by Shea. Currents \( i_c \), and \( i_e \), (both in amperes) and voltages \( v_c \), and \( v_e \), (both in volts) are maximum-signal peak values. Values are for two transistors except where noted otherwise.

1. D. C. Collector Power Input = \( (v_i)(i_c) \) \( 0.97 \) watts
2. Peak A. C. Power Output = \( (v_i)(2)(i_c) \)
3. Peak A. C. Driving Power = \( v_i \) watts
4. Load Resistance = \( \frac{4(v_o)(i_c)}{i_c} \) ohms (collector-to-collector)
5. Input Resistance per transistor = \( v_i \) ohms (base-to-base)

If \( i_1 \), (Figure 1) is taken as the peak value of the zero-signal collector current and \( i_1 \) as its maximum-signal peak value, the maximum-signal average value of the d. c. collector current for each transistor, as read with a D'Arsonval-type d. c. milliammeter in series with the collector, is:
Load Lines In Transistor Amplifier Design

Graphical constructions are of considerable aid in designing electron tube circuits. Load lines drawn across the plate current-voltage plate voltage family of curves yield circuit constants and important operating data. This relatively simple procedure eliminates many tedious calculations. Graphical constructions are equally useful in the design of transistor amplifiers. Similar advantages are obtained. In transistor work, load lines are constructed on the collector voltage, family of curves to determine operating point, load resistance, distortion, collector current swing, and base current values. The technique is similar in every respect to that employed in tube work.

In this discussion we will show by illustrative examples, how to use the load line techniques in transistor amplifier design.

Detailed Procedure

Figure 1 shows the circuit of a typical common-emitter, RC-coupled transistor amplifier stage. The common-emitter, also sometimes called the grounded-emitter, is widely used because it affords high power gain, high voltage gain, 180-degree phase reversal, similar to a tube, and good frequency response.

When the constants of this circuit, especially the load resistors (R₁ and R₂), collector supply voltage (Vₚ), and collector-emitter voltage (Vₑₑ), are chosen in such a way that the internal parameters of the transistor determine largely the operating point of the amplifier, inefficient operation, high distortion, or poor reliability usually result. By proper choice of the operating point with respect to the transistor characteristics and supply voltage, low-distortion clip-A performance easily is obtained safely within the transistor maximum ratings.

For graphical construction, the first requirement is to obtain a set of collector E-I curves, such as those in Figure 2, for the transistor chosen. For common-emitter operation, the family contains a separate curve for each of several typical bias-current values and a plot of collector-emitter voltage Vₑₑ versus collector current Iₑ. If such a set of curves is not available in the transistor manufacturer's literature, the operator must plot a set by making a series of common-emitter dc measurements on the particular type of transistor which is to be used in the amplifier. The procedure is to set the base bias current at a given level (zero base current is one such level) and to vary the collector current (supplied by a constant-current supply) which is observing the corresponding voltage between collector and emitter.

In the graphical construction, follow this procedure: (1) Label the points of maximum collector current and maximum collector voltage on the graph. Figure 2 is a family of curves for the Raytheon CK721 transistor. The maximum collector current, from the manufacturer's data sheet, therefore is 30 ma, as labelled in Figure 2, and the maximum collector voltage is 32 volts. (2) Draw across the curves a plot showing the current and voltage intersects for the maximum dissipation (Pₘ). In watts recommend by the manufacturer. In the case of the CK721, Pₘ = 33 milli-watts and is represented by the dashed line which bends across the collector family in Figure 2. At any point of intersection this line with abscissa and ordinates, the product X of the points for the dissipation curve, voltage, and current may be selected along the hori
mental axis and corresponding current values calculated ($I_e = P_{out}/V_{out}$), or current points may be selected along the vertical axis and corresponding voltage values calculated ($V_c = P_{out}/I_e$). The area of the graph below and to the left of this curve encloses all points which are within the dissipation rating of the transistor. All points in the area above and to the right of this curve represent overdrive and must be avoided. (c) Select the operating point of the transistor (that is, collector voltage and collector current) and the supply voltage. (d) Mark the operating point on the graph. Example, use the dot at the intersection of the 6-volt and 5-milliampere lines in Figure 2. (5) Construct a load line from the supply voltage point on the horizontal axis, through the operating point, to the vertical axis. This is the solid line in Figure 2. (6) Determine the required load impedance, $R_L$, by computing the slope of the load line ($Z_L = V_{out}/I_e$).

Several facts are evidence from an examination of Figure 2, a typical example of transistor load line application. (a) The operating point has been selected at $V_{cc} = 6$ v, $I_c = 5$ ma. (b) With a supply voltage of $12$ v, the load line extends from $-12$ to $-10$ ma. The load resistance accordingly is $R_L = (12 - 10)/0.002 = 1000$ ohms. (c) The entire load line is seen to be within the dissipation rating of the 2SK1, being 3 milliwatts lower than the constant dissipation curve at the point of closest proximity. (d) The base current level which will bias the transistor for the 5 ma. collector current at the operating point is seen to be 30 ma in this example. In the preceding example, the supply voltage was assumed fixed. The operating point was chosen, and the required load impedance determined from the slope of the load line. It is easily seen that different procedures might have been followed from combinations of known and unknown factors. For example, the operating point and load impedance are given, and the required supply voltage is to be determined, or the load resistance and supply voltage might be specified and the operating point placed at any satisfactory position along the resulting load line.

It is desirable that the collector signal voltage swing encompass as much of the collector characteristic as is possible. This will insure maximum output voltage. Thus, with the operating point set at 5 v, 5 ma in Figure 2, the collector signal voltage swing can be 20 ma base current curve and up to the load line to the 210 ma base current curve without encountering the severe bending at the left ends of these curves. However, it is obvious that the upper swing traverses a larger number of base current curves that the lower swing, because of the progressively closer spacing of the curves at the higher base current levels. The asymmetry of signal waveform due to this condition may be minimized by limiting the swing to the region of more nearly equal base current curve spacing, when low distortion is a more important objective than maximum output.

The collector current at the operating point is a function of the base bias current, $I_b$. Reference to Figure 2 shows that $I_c$ must be 0.9 microamperes when $R_b = 1250$ ohms and $R_c = 10$ volts. This base bias current may be obtained from a separate battery or from $V_b$ through a series dropping resistor, $R_b$, in Figure 1. Base bias also may be obtained from a voltage divider ($R_b/R_c$) operated from $V_{cc}$, and an emitter series resistor, $R_e$, in Figure 3. The required value of series resistor, $R_e$, in Figure 1, may be determined from the simple relationship $R_e = V_{cc}/I_e$, where $R_e$ is in ohms, $V_{cc}$ in volts, and $I_e$ in amperes. For the 0.9 microampere base current indicated in Figure 2, $R_e = 1250/0.9 = 1388.9$ ohms.

**Figure 3**

**Common-emitter amplifier with base-bias stabilization network**

**Figure 2**

**Collector to emitter voltage $V_{CE}$**

**Common-emitter collector family with load line construction (N-channel JFET: Transistor)**

**Figure 3**

**Common-emitter amplifier with base-bias stabilization network**

103
Simple, Inexpensive Geiger Counters

LIGHTWEIGHT, portable radioscopes are being actively sought by many organizations, and several successful models are now on the market. These offer more features for less money and are ready to be used immediately. The instrument may be designed for use in an automobile, a laboratory, or a portable unit. In any case, the instrument is designed to minimize the requirements for an automobile or laboratory setup. The instrument is designed to minimize the requirements for an automobile or laboratory setup.
Its threshold voltage is 280 v and its minimum plateau length 60 volts (plateau slope 30%/50 v). Its life is $5 \times 10^9$ counts. The counter tube may be operated in a probe on the end of a shielded cable, or it may be installed inside the instrument case, with a suitable opening, holes, or louvers for entry of radiant energy.

Remarkably high impedance, are unsatisfactory in this circuit because they provide no continuity for dc.

The 1886 is a glass-wall gamma ray tube. Its diameter is 0.4 inch and its over-all length, including its 1/8" tinned leads, is 4 inch. For more compact assembly or smaller-sized probes, Type 1886 counter tube may be substituted. The latter also has a diameter of 0.4 inch, but its over-all length (including 1/8" tinned leads) is only 2½ inches.

Circuit 1:

- Eight representative Geiger counter circuits are shown. These are discussed separately in the following paragraphs.

Circuit 1. This is the simplest possible arrangement. Here, a 1886 counter tube is connected in series with a miniature 200-volt battery (similar to Burgess U200 or RCA V5002), headphones, and a 1-megohm current-limiting resistor. A potentiometer switch allows the circuit to be disabled when not in use.

Each ionizing particle penetrating the counter tube causes a pulse of current to flow through the circuit, and this produces a click in the headphones. In order to complete the circuit, the headphones must be of a magnetic type and must have high impedance. A stepdown matching transformer would be required for low-impedance headphones. Crystal-type headphones, although possessing remarkably high impedance, are unsatisfactory in this circuit because they provide no continuity for dc.

Circuit 2:

The 300-volt Type 12600 battery measures 2 11/16" x 2 7/32" x 3 28/32".

Circuit 2. The sensitivity of the simple, battery-operated counter may be improved by employing a general purpose betagamma counter tube, such as Type 1880, with a 100-volt d-c supply. The high voltage is obtained from three miniature 300-volt batteries such as specified for single use in Circuit 1) connected in series.

Circuit 3:

The 1886 and 1U5 circuit diagram is shown.
Because the 300-volt potential constitutes a serious shock hazard, we do not recommend connecting the headphones in series with the other elements in the way this was done in Circuit 1. Instead, high-impedance headphones are capacitance-coupled to the circuit through a 300-volt 1500-microfarad capacitor. Crystal-type headphones can be used to advantage in this circuit.

The 1865 is an aluminum wall tube having an A-152 coaxial base. Its diameter is 5/16 inch, and its overall length, including the coaxial base, is 4-3/4 inches.

Because of the higher dc voltage employed in Circuit 2, particular care must be exercised during construction of the instrument to insulate the various parts of the assembly.

To prevent leakage, use only high-grade non-hygrosopic insulating materials and keep all circuit points separated as widely as possible.

Circuit 5. In this arrangement, a pentode amplifier has been added to the rudimentary circuit (Circuit 1) to increase sensitivity and to provide louder clicks. Pulses from the 1865 counter tube are transmitted to the grid of the 115 amplifier through coupling capacitor C1.

The midsize 300-volt battery, B1, supplies both the counter tube and the plate and screen of the amplifier. The second battery, B2, is a 1-1/2 volt Size-C flashlight cell. The spot ON-OFF switch is connected to open both battery circuits simultaneously when in its OFF position. Maximum efficiency will be obtained with high-impedance headphones. Crystal-type phones are recommended.

The shorter Type 1865 counter tube also may be employed in this circuit. Circuit 4. This is a variation of Circuit 3. Here, the plate and screen of the 115 tube have been disconnect- ed from the important 300-volt battery and are supplied by the separate, midsize 675-volt battery, B3. In all other respects, Circuit 4 is identical to Circuit 3.

This circuit change is important when the instrument is to be used regularly over protracted periods. Under such circumstances, removal of the tube drain from the 300-volt battery and the consequent longer life of this battery sufficiently offsets the additional cost, complication, and weight of the additional battery, B3, to warrant the modification.

The spot switch interrupts all three battery circuits when thrown to its OFF position.

Circuit 5. This is an interesting 300-volt Geiger counter which is powered from a single 1-1/2 volt Size-D flashlight cell. The high dc voltage required for operation of the 1865 counter tube is obtained by charging a capacitor through a high turns-ratio stepup transformer.

The transformer may be any small universal replacement-type output unit, such as Merit A2960. Terminals A and B represent the entire primary winding of this transformer. (Do not use the center tap.) Terminals C and D are two of the low impedance connector pins selected from the transformer connection chart for a turns ratio between CD and AB of 300 to 1, or higher.
The split pushbutton switch normally rests in position a. This connects the headphones (plugged into the insulated phone jack) to the low-impedance winding of the transformer. When the switch is depressed and released momentarily, making a quick make-and-break connection in position b, a high-voltage pulse is induced across the high-tension winding Ab. This pulse fires the 1HM tube and charges the 0.1- microfarad capacitor. Several repeated operations of the pushbutton switch will charge this capacitor fully. With the switch resting in position a, the capacitor cannot discharge through the circuit because of the insulating property of the 1HM when it is not fired. The capacitor accordingly retains its charge and serves as a 900-volt source. When the counter tube is penetrated by a radioactive ionizing particle, however, it fires and induces a voltage across winding C2 and this produces a click in the headphones.

The length of time the capacitor will hold its charge depends upon the leakage characteristics of the capacitor and how well the circuit is protected from ambient humidity. This will vary from 5 to 30 minutes in most cases. As the charge leaks away, because of either leakage or rapid counting, the capacitor may be recharged readily by quickly pulsing the pushbutton switch several times. Because of the nature of this circuit, very small size may be attained in the completed instrument. An added advantage is the fact that the headphones, being inductively coupled, are not included directly in the high-voltage circuit and any possibility of shock, ground, or short circuit consequent is removed.

Circuit 6. The high potential of 900 volts is obtained in this circuit also by pushbutton pulsing a 1½-volt cell through a backward-connected universal output transformer to

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CIRCUIT 6

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- **PUSHBUTTON**
- **1885**
- **HIGH-IMPEDANCE MAGNETIC HEADPHONES**
- **0.088" NEEDLE GAP**
- **S1**
- **L1**
- **R1**
- **C1**
- **B1**
- **10 MEG 1/2 W**
- **0.05 μF 1000 V**
- **10 MEG 1/2 W**
- **B2 22½V**
- **ON-OFF**

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CIRCUIT 7

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- **HIGH VOLTAGE SELENIUM CARTRIDGES**
- **SE1**
- **SE2**
- **50μF 150 V, ELECT.**
- **5 MEG 1/2 W**
- **C1**
- **0.1μF, 600 V, TUBULAR**
- **5V 1/2 W**
- **C2**
- **SE3**
- **1885**
- **3A5**
- **R5 18K 1/2 W**
- **C3**
- **R4 250 μF 10 W**
- **C4**
- **R6 0.1M 400 V TUBULAR**
- **C5**
- **ON-OFF**
- **VIBRATOR UNIT**
- **5641**
- **CHASSIS**
- **M**
- **IN34**
- **CIRCUIT PHASE J0CK (HIGH-IMPEDANCE MAGNETIC PHONES ONLY)**
- **R8 5 MEG**
- **G5 0–50 DC μA**
- **CALIBRATION CONTROL**

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107
charge a capacitor. Here, however, terminals 1 and 2 of the transformer are the lugs (selected from the transformer connections chart) which supply the lowest impedance. This permits the highest output ratio. 3 and 4 represent the entire primary winding of the transformer. (Do not use the center tap.)

A small needle-type spark gap is connected between the 1B85 tube and the high-voltage winding of the transformer. A discharge takes place across this gap when the pushbutton (S1) is depressed and released, and the capacitor is charged. Several, successive, rapid pulsations of the pushbutton will charge the capacitor fully.

As the 1B85 tube fires under the influence of penetrating radioactive particles, the resulting pulses are presented to a grid of the 2N4 amplifier tube, and the amplified pulses are delivered to the headphones.

The 12-volt battery, B1, supplies both the transformer and the 2N4 filament. The 22½-volt battery, B2, supplies only the plate and screen of the 2N4. For light-duty application, B2 may be a single Size-D flashlight cell and B1 a hearing aid battery. For more exacting work, B2 should be made up of two or more Size-D cells connected in parallel, and B1 should be a midget 22½-volt radio-type B-battery.

As in Circuit 5, the length of time that the capacitor will remain charged will depend upon ambient humidity and capacitor leakage and usually will be between 5 and 30 minutes. An occasional succession of rapid pulsings of the pushbutton switch, S1, will restore the charge.

Circuit 7. The Geiger counter has a self-contained, miniature, vibrator-type power supply operated from a single 12-volt Size-D flashlight cell. The 2N4 tube supplies the 1B85 counter tube and the 2N4 triode amplifier tube. Meter readings, as well as headphone signals, are obtained with this circuit.

The miniature vibrator unit is a Model 10MV, a product of Precise Measurements Co., Brooklyn, N. Y. It consists of a vibrator integral with a high-voltage transformer. The spark-suppressing capacitor, C2, is self-contained.

The high-voltage a-c output of the transformer is rectified by two miniature high-voltage selenium rectifiers. One of these, SE1, is polarized to supply positive output for the 2N4 tube. The other, SE2, is polarized for negative output for the 1B85. The 50-microfarad, 50-volt electrolytic capacitor, C3, filters the 2N4 voltage, and its normal leakage holds this voltage approximately to 80 volts at 200 microamperes. The Victron Type 5601 regulator tube holds the counter tube voltage to a constant 900-volt level. Capacitors C2 and C3 and resistor R1 form the high-voltage filter.

The 2N4 tube provides a 2-stage RC-coupled amplifier. The output triode is capacitance-coupled, through C4, to a rectifier-type voltmeter calibrated by the 1NO4 diode and the 0.5-microammeter, M.

The meter deflection is proportioned to the number of pulses per unit time arriving from the counter tube. Its scale may be calibrated at various settings of potentiometer B1 (which might also be a step-type attenuator) with the aid of a series of calibrated radioactive samples held close to the 1B85 tube. Magnetic-type headphones inserted into the phone jack will provide aural indications. Since this jack is of the closed-circuit type, it will restore the plate circuit of the second triode automatically when the headphone plug is extracted.
Circuit 8. This circuit employs a 3AS dual triode in a cathode-coupled one-shot multivibrator circuit. The multivibrator is triggered by pulses from the 1885 counter tube. Each pulse switches the multivibrator on and off, causing a single pulse to be delivered to the metering circuit. Capacitors C0, C1, and C2, switched into the circuit, alter the pulse duration and repetition rate of the multivibrator and thus change the meter range. The microammeter scale accordingly may be calibrated in counts per unit time, milliroentgens per hour, or similar units. For this purpose, a series of calibrated radioactive samples may be employed.

Like a VSI voltmeter, the meter is set initially to zero in the absence of any input signal, by adjustment of the potentiometer R2. An aural indication is obtainable from magnetic-type headphones plugged into the closed-circuit jack.

This multivibrator-type Geiger counter circuit is adapted from a similar one developed by Friedland. (See Radiological Monitoring, Stephen E. Friedland, QST, June 1951, p. 29.)

The 900-volt potential for the counter tube is developed by a special power supply based upon a neon-bulb relaxation oscillator. Operated from the same 4⅛-volt battery, B6, that furnishes the 3AS plate voltage, the relaxation oscillator consists of rectifier R6, capacitor C2, and the NE-2 neon bulb. The sawtooth voltage developed by this oscillator is applied to the grid of the 1885 tube. A choke coil, consisting of a miniature "Oncor" transformer, T, with its primary and secondary connected together in series-adding, is connected in series with the 1885 plate. The rapid fall of current through this choke, due to the decay ("flyback") of the sawtooth wave, induces a high voltage across the choke. This high voltage then is rectified by the high-voltage selenium cadmium, filtered by R8 and C8, and applied to the 1885 tube. The high-voltage dc may be set exactly to the required level of 900 volts (preferably with the aid of an electrostatic voltmeter) by adjustment of potentiometer R2.

A separate 14-volt cell, B3, is required for the 3AS filament, since the cathode-coupled circuit in which this tube is used requires that the filament "float" 3000 ohms above ground. B3 and B6 are size-D flashlight cells. B6 is a radio-type 4⅛-volt B-battery.

Recent Trends in Single-Sideband Communication

ROM the days of the earliest radio-telephone transmissions, radio engineers and experimenters have been impressed with the fact that a conventional amplitude-modulated signal contains all its necessary intelligence in one of its sidebands. Yet associated with this sideband is an identical additional sideband and a carrier component, so the total transmitted power amounts to six times that of a single sideband, which contains all the necessary intelligence. At the time, the conventional amplitude-modulated signal takes up twice as much precious frequency spectrum bandwidth as a single sideband.

It has been found that certain types of fading, common with amplitude-modulated signals, are either eliminated or greatly reduced in single-sideband transmission. This is particularly true of selective fading which results from propagation phase variations between carrier and sidebands or between the two sidebands themselves. Elimination of the heavy and expensive high-level audio-frequency components of the conventional a-m transmitter is a powerful argument in favor of the single-sideband system.

Radio transmission with suppressed carrier and a single sideband has been used almost as long as the conventional double-sideband amplitude-modulated type. It was employed for transatlantic radiotelephony at low frequencies as long ago as the early nineteen twenties. Because at that time the filter method was the only one generally applied, the importance of heterodyning, instability of oscillator components and other factors limited practical use of such a system to the lower radio frequencies. In the last few years, development of high grade components and new techniques of circuitry have made single-sideband-suppress-carrier (ssc) communication at higher frequencies more attractive. Improved stability and selectivity in
receivers and receiving systems have also been helpful in this development.

The Filter Method of Generating SSB Signals

Two main methods of generating single-sideband signals are generally employed: the filter method and the phasing method. The filter method is the older of the two, and in principle is simpler and more direct. This principle is illustrated in Fig. 1. The carrier is removed from the composite modulated signal by means of a balanced modulator. Then a filter with a sharp frequency characteristic, connected at the output circuit of the balanced modulator, removes the unwanted sideband.

In the filter method, the carrier must be generated at a relatively low radio frequency. If the radio frequency is too high, the percentage frequency spacing between the two sidebands is too small. Filters sharp enough to isolate the desired sideband would introduce too much phase distortion. Some amateur installa-
tions are using generated rf carriers as high as nearly 500 kc for the filter method. Crystal lattice filters are used. However, in commercial prac-
tice it is common to keep the generated carrier between 20 kc and 100 kc to ensure high filter efficiency and best attenuation of the unwanted sideband.

The single-sideband signal from the balanced modulator and filter is then transposed to the desired higher operating frequency by heterodyning with higher frequency carriers in one or more successive mixers. If the operating frequency is many times that of the carrier of the sideband generator (as is ordinarily the case in high-frequency transmitters of the filter type), more than one stage of mixing is necessary. This is because when the mixer output frequency is very much higher than its input frequency, and the frequencies of the sum and difference heterodyne products are close, it is difficult to attenuate the unwanted heterodyne. The necessity for one or more mixing stages and lack of flexibility in switching from one sideband to the other are disadvantages of the filter method as applied to the higher communications frequencies.

The Phasing Method

In most cases, the need for mixer stages can be eliminated by use of the phasing method of single-sideband generation. In this method, the rf carrier can be generated at the transmitter's output operating frequency. The principle is illustrated in Fig. 2. The mathematical expressions for the various signals, assuming a single-tone modulation signal, are given alongside the respective paths. Both the intermediate-frequency modulating signal and the carrier are divided, each into two components in quadrature. As shown, each phase component of the rf signal is combined with one phase component of the carrier in the balanced modulator. In each case, the balanced modulator removes the carrier, leaving the two sidebands in the output from each modulator. In the example given, assuming the use of a single modulation frequency, there would be just one single-frequency side-component on either side of the carrier frequency (but no carrier) in each case.

After the balanced modulators, their outputs are combined. The phase relations of the four sidebands (or single-frequency components) is such that one sideband is canceled while the other is reinforced. In the example illustrated, the carrier component which was shifted 90 degrees is modulated by the a-f component which was shifted 90 degrees. As can be seen by the mathematical expressions, combination in such phases results in cancelling of the high frequency sideband and reinforcement of the low frequency sideband. The phase of either the a-f or the carrier component can be reversed to provide high frequency sideband output instead of low frequency sideband output.

One of the critical sections of the phasing type transmitter is the audio phase shifter, which must provide two a-f signals 90 degrees apart at all frequencies in the modulation range. For the normal voicecommunication (frequency range, a simple rc network will accomplish this. A typical phase shift network for this purpose is illustrated in Fig. 3. One of the problems connected with such a circuit, of course, is the odd values required for most of the components. Special, non-standard components would have to be obtained.

However, a big step forward in making the phasing type of transmitter more practical was the provision by several manufacturers of complete phase shifting networks in very compact form. One of these standard commercially-available networks is completely enclosed in a metal tube envelope, of the same size as a 6sl
tube, and plugs into an octal socket. It provides a 90-degree phase shift constant to within ± 1.5 degrees over an audio frequency range of 300 to 3,000 cps, adequate for voice communication purposes.

**Linear Amplifiers**

Use of one communication was spurred renewed attention to linear amplifier design. Two factors are important in these amplifiers: (1) low distortion, and (2) maximum power gain. Stability, which is very important, is interrelated with distortion, so is not classified separately.

Realization of the full advantages of the limited bandwidth of single-sideband transmission requires that distortion be well controlled. This is especially true in services in which each sideband is used for transmission of different intelligence, in which case distortion modifies itself in spurious and noise effects between channels. Thus, low distortion can be considered even more important in all linear amplifiers than in the conventional carrier systems.

On the other hand, high power gain is also an important consideration. Single-sideband generators must be operated at relatively low levels. Thus, if linear amplifiers which follow do not have a high power gain, a large number of stages must be used and this is uneconomical and leads to unreasonable maintenance and adjustment requirements.

Under ordinary conditions, very low distortion and high power gain are not compatible. High power gain is seldom obtained with best linearity and low distortion. It is the reconciliation of these two factors which is the objective of modern linear amplifier development.

The conventional grounded-cathode linear amplifier exemplifies the high-power-gain type, especially those circuits employing tetrodes and beam tetrodes. However, in the latter, optimum power gain is not consistent with minimum distortion. When the screen voltage is raised to a high enough level to minimize grid drive requirements, the static plate current becomes so high that the plate dissipation is likely to become excessive. On the other hand, when the static plate current is kept low, distortion is introduced. One way to maintain the low grid drive requirement and still keep static plate current limited is to use negative feedback.

One popular way to introduce feedback to improve linearity and stability is to use the grounded grid circuit. Although this does provide excellent linearity, the power gain is low.

**Two-Stage Feedback**

A somewhat different arrangement for obtaining high power gain with low distortion is the two-stage amplifier. The first stage is operated at a relatively low level. The output is then amplified by a second, higher power-gain stage, with the output fed back to the grids. The output of this stage is then amplified by a third stage, with the output fed back to the grids of the second stage. This arrangement is sometimes referred to as a "push-pull" amplifier. The feedback is usually provided by means of a coupling capacitor, which is coupled between the grids of the two stages.

Even though tetrodes are used, neutralization is provided; this is desirable to minimize single-stage instabilities. Less circuit "swinging" is then required.

**SSB Reception**

The most efficient method of reception of SSB signals is the phasing arrangement, which is the reverse of the transmission method illustrated in Fig. 3. A block diagram of the front end of the phasing-type receiver is shown in Fig. 4. In its applications in which a "pilot" carrier level is transmitted, automatic frequency control is employed in conjunction with the local oscillator. Otherwise, a suitable vernier adjustment of oscillator frequency must be available to facilitate manual control.

General improvement in component and material quality have contributed greatly in bringing set receivers to a relatively high state of development. The narrow pass band
The Citizens Radio Service

The Citizens Radio Service was inaugurated by the Federal Communications Commission in 1949. It was designed to provide a low-cost communication service for amateur radio enthusiasts. The service was later expanded to include a range of communications activities, including emergency communications. The Citizens Radio Service is regulated by the Federal Communications Commission (FCC) and operates on frequencies that are not designated for commercial use. The service is open to anyone who obtains a valid license, and it provides a way for individuals to communicate with each other using radio equipment. The Citizens Radio Service is particularly popular among hikers, hunters, and other outdoor enthusiasts who need reliable communication in remote areas. The service is also used by organizations and governments for emergency communications. The Citizens Radio Service is a testament to the ingenuity and spirit of innovation that characterizes American engineering and technology.
of the mid-band frequency (465 Mc.) and all operation must be confined to this band. The maximum power input to the final tube or tubes in the transmitter for the three subdivisions of the band is also shown in Fig. 2.

The technical requirements of the Class A license effectively exclude all but carefully engineered, crystal controlled transmitters. No other known method of transmitter frequency control will dependably meet the 0.05% tolerance specified. Since low-drift quartz crystal oscillators are available only at relatively low frequencies, it is necessary to use a multi-stage transmitter to multiply the crystal frequency to the required Citizens band frequency. The usual extent of such a crystal controlled transmitter is illustrated in block form in Fig. 3. It is common practice to accomplish the frequency multiplication in low power stages and use the output of these to drive a high-gain power amplifier at the output frequency.

It is evident from Fig. 3 that the Class A requirements may be met equally well by stations operating at fixed locations, or in the more elaborate mobile installations. The inherent complexity of the high stability equipment and the accompanying heavy power drain make the hand-portable "walkie-talkie" type of operation impractical. The same equipment is necessary for operation near the edges of the citizens band to avoid interference with important commercial services on adjacent frequencies. The recent allocation of UHF television channels starting at 475 Mc. has made the good local Citizens band stations in the Citizens Radio Service even more imperative. For this reason equipment meeting FCC approval must be as "fool-proof" as possible. Since it is intended for use by non-technical personnel, no control affecting the frequency of transmission should be accessible from the exterior of the equipment housing.

It is a Class B license which is intended to authorize the operation of generally simplified and possibly less expensive transmission equipment which might be carried on the person in the manner of the war-time "walkie-talkie" or "handie-talkie" sets. This mode of operation has the greatest appeal to the public and probably the greatest potential field of application. The technical requirements of this license may be met by relatively simple, lightweight set of the "transceiver" type in which the functions of transmitting and receiving are performed by the same tubes. Fig. 4 shows the minimum equipment requirement for such a station in block form. The transmitter is a self-excited oscillator, modulated by one or two audio amplifier stages driven by a high-output microphone. During receiving, the bias and plate voltage on the oscillator tube are switched to convert it to a superregenerative detector. The antenna remains coupled in the same manner, and the audio section is used to amplify the detected signal and drive a small speaker or headphones. The complete transition from "transmit" to "receive" is accomplished by a switch or relay. Although equipment of the transceiver type has been approved by the FCC for use in the citizens band, very careful engineering is required to comply with the Class B regulations. This difficulty is due to several inherent limitations in the performance of modulated self-excited oscillators and transceivers in general. These limitations may be listed as follows:

(a) Transmitter frequency instability. The inherent frequency stability of most self-excited oscillators in the UHF range is very poor. The frequency is effected by changes in applied voltage, capacity changes caused by thermal expansion of vacuum tube elements, changes in ambient temperature and humidity, capacity variations due to proximity effects, antenna loading variations, and mechanical vibration. Voltage fluctuations due to plate modulation alone may cause frequency modulation amounting to several hundred kilocycles. The combined result of such effects is to make it exceedingly difficult to design an oscillator at 465 Mc., which will meet the prescribed tolerances in the hands of the public.

(b) Receiver radiation. The great simplicity of the transceiver results from the dual use of the tubes. However, when the transmitting tube is used as a superregenerative detector coupled directly to the antenna, a signal which is pulsed at the frequency of the quenching is radiated. This signal is a potential source of interference to other citizens radio stations and adjacent services. Receiver radiation may be prevented by the addition of an r-f stage between the antenna and the "superregen" detector, but the presence of this stage complicates the function of transmit-receive switching considerably.
(c) Transmit-receive frequency difference. Due to the change in operating voltage, there is a considerable difference between the frequency of transmission and reception in most transceivers. This discrepancy in frequency has the effect of causing two stations in communication to drift around the band unless the transmitter of each is returned to a given frequency before every transmission.

Nevertheless, the transceiver presents a very inviting approach to the problem of simple and economical equipment for the Citizens Radio Service, and the solution of its limitations is a challenge to equipment designers.

Propagation Characteristics

Because the citizens band allocation is in the UHF portion of the radio spectrum, communication is usually limited to virtual line-of-site distances. Stations using a few watts input, as is characteristic of the "man-portable" kind, are able to communicate for distances ranging from a few blocks in very populous city areas, to several miles in rural districts having fairly flat terrain. Over truly optical paths, such as might exist between stations operating at elevated locations, appreciable distances may be spanned, even with very low power. The greatest record for amateur communication in the 420-450 Mc. band is 262 miles.

Waves at these frequencies are attenuated considerably by dense foliage. Antennas should, therefore, be above tree-top level wherever possible. Because of the portable applications of the service, vertical antenna polarization will be favored for its omnidirectional properties.

There are also very pronounced "shadowing" effects behind hills, tall buildings, and other large obstructions at these frequencies. In many cases, however, reception at such sites may be made possible by indirect path propagation of the type illustrated in Fig. 5. By means, radio waves reflected by other large obstacles may reach the receiver by indirect routes.

Citizens stations using the high-performance equipment specified for the Class A license, and the greater power permitted under these conditions, should enjoy considerably greater dependability of communication. The use of very stable transmitters will permit relatively narrow-band, high gain superheterodyne receivers to be employed. The resulting improvement in receiver sensitivity should enable this type of station to communicate over distances several times greater than those covered by Class B stations. High-gain directional antennas may be employed by stations at fixed locations.

Commercial Aspects

The Citizens Radio Service represents a new field of endeavor for the manufacturer and radio service man. Since the licensing procedure for the service is based primarily upon the availability of commercially manufactured equipment, a large potential market exists for type-approved sets which can be economically produced. The high cost of satisfactory equipment for two-way radiotelephone communication is at present a major drawback in the expansion of this service.

In a similar manner, citizens radio provides a lucrative field for the service man who equips himself with a commercial radiotelephone license so that he may legally engage in installation and maintenance of citizens radio sets. As the service grows and gains in popularity it should be possible for an enterprising technician to build up a large clientele of citizens band licensees. Since the service man will have to assume responsibility for the proper functioning of the equipment under his care, it will be necessary for him to acquire some specialized test equipment, such as a precision UHF wavemeter, for this purpose.

ELEMENTARY BINARY ARITHMETIC

With the growing presence of digital electronic computers among us, more and more radio technicians are beginning to bear oblique-ly about the binary number system and wonder why they have learned so little. There is good excuse for the perplexity, since surprisingly little has appeared on the subject in the books and magazines customarily read by radio men. Many technicians who have prided themselves on being reasonably well-grounded have thumbed through mathematics textbooks, old and new,
and found no reference whatever to the binary system!

A glance into the dictionary reveals the word binary to mean "characteristic of two things or parts." From this, we may infer, correctly, that binary arithmetic is in some way associated with the figure 2. Indeed, the binary system uses only two digits. Now, let us see how this differs from the method of counting we have employed most of our lives.

Our old standby is the decimal system. Its base is 10 and its digits are 0, 1, 2, 3, 4, 5, 6, 7, 8, and 9. This is very handy because we have ten fingers on which to count. In our civilization, we've gotten along famously with the base 10. It is possible to express any number by the proper combination of the digits 0 to 9. However, when we attempt to set up some forms of electrical counting equipment in strict accordance with the decimal system, we find ourselves in need of a multitude of components.

Here, the binary method comes to our rescue. It is a base 2 system and requires only two digits: 0 and 1. In the binary system, all numbers can be expressed by combinations of zeros and ones. Just why should this be handier than the decimal system? Simply because it is an easy matter to express the binary digits themselves with a simple electrical device which is either ON (1) or OFF (0). Thus, an open switch or relay denotes zero, while a closed switch signifies 1. The same is true of a tube conducting or cutting off, a crystal diode conducting or blocking, a neon lamp ignited or extinguished, etc. etc. Voltage or current likewise can denote 1 when high or positive, and zero when low, negative, or off. The binary system operates with fewer and simpler components.

Although a certain piece of equipment, such as a counter, might operate by the binary method, it will can be made to give indications (such as total count) in the easily-recon- nited decimal notation.

In explaining the elements of binary arithmetic in this article, frequent comparisons will be made with the decimal system for the sake of clarity or proof.

**Binary Ciphers of the System**

Suppose that you have four separate on-off components (switches, tubes, etc.), each of which is assumed to indicate zero when OFF and 1 when ON. Table 1 shows how the two states of these four devices can be employed to express various decimal numbers.

In order better to understand this table, let us consider the basic rules of binary addition which may be stated as follows: 0 + 0 = 0; 0 + 1 = 1; 1 + 0 = 1; and 1 + 1 = 0. This last sum means simply that every time 1 is added to 1, we write down zero and carry 1 to the next column to the left. An illustration will serve to clarify binary addition. For example, from Table 1 add 0111 (binary 7) and 0011 (binary 3):

<table>
<thead>
<tr>
<th>BINARY</th>
<th>DECIMAL</th>
</tr>
</thead>
<tbody>
<tr>
<td>1000</td>
<td>8</td>
</tr>
</tbody>
</table>

First, the two 1's in the right-hand column are added. This equals 0, so we write 0 and carry 1 to the next column to the left. This 1 must be added to 1 already in that column. Again, this equals 1, so we write 0 and carry 1 to the next column to the left. Adding this carried 1 to the 1 already in that column gives another 1. Since we have added 1 to 1, we are carried 1 to the left-most column. Now, this 1 is added to the zero in that column, giving 1 which is written. The answer is 1000, which by reference to Table 1 is found to be binary 8.

A careful examination of Table 1 now reveals that each higher binary number is obtained by adding binary 1 (0001) to the preceding number. Try this out by starting with 1010 (binary 10) and successively adding 0001 (binary 1). You will obtain 1011 for 11, 1100 for 12, 1110 for 14, and 1111 for 15. If you make another addition you will obtain 10000 (binary 16) which requires five on-off devices for its expression — and we agree at the beginning that we have only four. So binary 15 is as high as a 4-device system will count. However, the economy and efficiency of the system is realized when it is considered that only four elements are needed to display from 0 to 15 events.

Any number may be expressed in the binary system by choice and position of the two digits repeated as coefficients of powers of 2, just as any number can be expressed in the decimal system by choice and position of the ten digits of that system as coefficients of powers of 10. For example: The decimal number 2564 means $2 \times 10^3 + 5 \times 10^2 + 4 \times 10^1 + 8 \times 10^0$. Similarly, the binary number 011010 means $0 \times 2^5 + 1 \times 2^4 + 1 \times 2^3 + 0 \times 2^2 + 1 \times 2^1 + 0 \times 2^0$. Table 1 lists the powers of 2 up to 25, and you can obtain from this Table the decimal numbers corresponding to the powers of 2 given in the preceding example. Adding these discloses that 011010 equals 26:

0 \times 2^5 = 0
+ 1 \times 2^4 = 16
+ 1 \times 2^3 = 8
+ 0 \times 2^2 = 0
+ 1 \times 2^1 = 2
+ 0 \times 2^0 = 0

011010 = 26

You can prove this sum by returning to Table 1 and adding binary 1 (0001) successively to binary 10 until you reach 011010 which you will find equal to 26.

A binary point is used in binary notation just as a decimal point is used in decimal notation. An example is 0.1010101 with six digits on
the left and two on the right of the binary point, although the digits might increase in number without limit. We have seen already that the digits on the left of the point are coefficients of increasing negative powers of 2 with 2° adjacent to the binary point. The digits on the right are coefficients of increasing positive powers of 2 with 2° adjacent to the point. The example just given (100101.01) becomes:

\[
\begin{align*}
1 \times 2^2 &= 32 \\
+ 0 \times 2^1 &= 0 \\
+ 0 \times 2^0 &= 0 \\
+ 1 \times 2^{-1} &= 4 \\
+ 0 \times 2^{-3} &= 0 \\
+ 1 \times 2^{-4} &= 1 \\
+ 0 \times 2^{-5} &= 0 \\
+ 1 \times 2^{-6} &= 0.25
\end{align*}
\]

\[100101.01 = 37.25\]

**Binary Multiplication**

Binary multiplication is carried out in very much the same manner as decimal multiplication, obtaining partial products in the conventional manner, but adding the latter in binary fashion. In binary multiplication, \(0 \times 0 = 0. 0 \times 1 = 0, 1 \times 0 = 0\), and \(1 \times 1 = 1\).

As an example, multiply 0101 (binary 5) by 0010 (binary 2):

\[
\begin{align*}
0101 \times 0010 &= 00001010 \\
0000 &\quad \times 2
\end{align*}
\]

\[0001010 = 1010 = 10\]

**Binary Division**

Binary division is carried out in a manner similar to decimal division, as the following example will show. Divide 1001 (binary 9) by 0100 (binary 4):

\[
\begin{align*}
10.01 &= 9 \\
0100 \div 0100 &= 100 \\
000100 &
\end{align*}
\]

The quotient 10.01 =

\[
\begin{align*}
0 &\times 2^3 = 0 \\
0 &\times 2^2 = 0 \\
0 &\times 2^1 = 0 \\
+ 1 &\times 2^0 = 0.25
\end{align*}
\]

\[10.01 = 2.25\]

**Binary Addition**

The addition of positive numbers in the binary system already has been explained in the preceding paragraphs. While the addition of two numbers has been given in each illustration, the system is by no means restricted to 2 number groups. Any series of binary numbers can be summed.

The only remaining case is the addition of a positive and a negative number. Consider, for example, the addition of 0101 and -0010.

<table>
<thead>
<tr>
<th>DECIMAL</th>
<th>BINARY</th>
</tr>
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<tbody>
<tr>
<td>5</td>
<td>0101</td>
</tr>
<tr>
<td>-2</td>
<td>-010</td>
</tr>
</tbody>
</table>

The technique is to change the sign of the negative number, then complement this number, and add the result to the positive number. To complement the number, change each of its 1's to 0's and each of its 0's to 1's and add 1. Thus, -0010 becomes:

\[0011\]

Here, the left-most digit in the answer is discarded. If it is 1, the sign of the answer is positive, as in the above case. The answer thus is + 0011, or binary 3 which satisfies the condition of 52 = 3.

If the left-most digit is zero, the sign of the answer is negative and the result must be re-complemented (the same process as the original complementing) to give the correct answer. This always happens when a negative number is added to a smaller positive number. Thus: Add -1000 (binary 8) to 0011 (binary 3):

\[0011 + (-1000) = 0011 \\
+ 1000 = 01011 \\
-1000 = \text{complemented}
\]

Dropping the left-most 0 in the answer (which merely indicates the negative sign), and re-complementing changes 0011 to -0101 (binary 5), which is the correct answer. \(-8 - 3 = -5\).

**Binary Subtraction**

Subtraction is the same as the addition of positive and negative binary numbers, as just described. For example: Subtract 0100 (binary 4) from 1010 (binary 10). 1010-0100 becomes 1010-1100 when the negative number (subtrahend) is complemented and its sign changed. This equals 1010. The left-most digit, being a ‘1’, indicates that the sign of the answer is positive and is discarded, making the answer +0110 (binary 6) which is correct.

116
Conclusion

After studying the rudiments of binary arithmetic presented here, the reader should be able, by setting up for himself a number of practice examples for drill, to acquire considerable proficiency in manipulating this invaluable new tool. A good working knowledge of the binary system is essential to comprehending the operation of digital electronic computers and of other instruments such as counters, which utilize the digital techniques.

<table>
<thead>
<tr>
<th>DECIMAL NUMBER</th>
<th>BINARY NUMBER</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>0000</td>
</tr>
<tr>
<td>1</td>
<td>0001</td>
</tr>
<tr>
<td>2</td>
<td>0010</td>
</tr>
<tr>
<td>3</td>
<td>0011</td>
</tr>
<tr>
<td>4</td>
<td>0100</td>
</tr>
<tr>
<td>5</td>
<td>0101</td>
</tr>
<tr>
<td>6</td>
<td>0110</td>
</tr>
<tr>
<td>7</td>
<td>0111</td>
</tr>
<tr>
<td>8</td>
<td>1000</td>
</tr>
<tr>
<td>9</td>
<td>1001</td>
</tr>
<tr>
<td>10</td>
<td>1010</td>
</tr>
</tbody>
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TABLE II

<table>
<thead>
<tr>
<th>20</th>
<th>1200</th>
<th>8192</th>
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<tbody>
<tr>
<td>21</td>
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<td>23</td>
<td>1208</td>
<td>65536</td>
</tr>
<tr>
<td>24</td>
<td>1216</td>
<td>131072</td>
</tr>
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<td>25</td>
<td>1232</td>
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</tr>
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<td>26</td>
<td>1264</td>
<td>524288</td>
</tr>
<tr>
<td>27</td>
<td>128</td>
<td>1 048 576</td>
</tr>
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<td>28</td>
<td>1256</td>
<td>2 097 152</td>
</tr>
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<td>1512</td>
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<td>30</td>
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<td>16 777 216</td>
</tr>
<tr>
<td>32</td>
<td>1806</td>
<td>33 592 432</td>
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</tbody>
</table>

Positive Powers of 2.
Single-sideband communication—continued
110
phasing method, 110
principle of, 108, 110
S/N ratio, 111, 112
two-way reception, 111, 112
Stagger-tuned cols, 26, 27
Stagger tuning, 9, 10

T
Television finals test, 10
Television interference filters, 9-11
Thermistor, 67
Thyrite resistors, 44-45
circuits, 45, 46
Tracking oscillator, 24-27
circuit, 25
error curve, 25
Transistors, 90-104
amplifying crystal, 90-95
base-collector curve, 108
basic circuit, 35
characteristics, 94, 95
class-B amplifier, 98-102
voltage-modulator curve, 100
junction, 94-99
load line, 102-104
maximum power output, 104
Transmission line runs, 31
Transmitter filter, low pass, 31
Tuned circuit feedback, 13, 19
Tuned grid feedback, 13
Tuned plate feedback, 16
Tuned plate tuned grid circuit, 17

TVI, 9-11
cause of, 9
reduction at receiver, 11
reduction at transmitter, 9, 11
TV receiver filter, high pass, 11

U
Unipolar circuit, 29-30
bipolar tank, 30, 31
used in TV, 30, 31
Using standard time and frequency broadcasts, 85-87
accuracy of transmitters, 86
audio-frequency comparisons, 87
calibration, low-frequency R.F., 86, 87
carry frequency checks, 86
methods of comparison, 86

V
VHF and UHF oscillators, 55-58
circuit construction, 55, 56
predistortion, 27, 28
area, 38, 39
transmission line tank, 31
type of circuits used, 30, 31
Video 1F, amplifiers, design of, 4-6
bandwidth, 4
response curve, 5
Voltmeter, D.C., 68

W
Wiring and cabling, 53
Wiring techniques, 44-54
WWV and WWVII, 83, 86