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In This Issue...

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Editorial . . .

IN RETROSPECT—A YEAR OF PROGRESS

by John E. Remich
Manager, Technical Department

As another year comes to a close, it is natural to pause momentarily to review the progress that has been made during the year. Progress in all scientific fields has been great; however, the electronics field is the one which is of common interest to the BULLETIN readers.

During the past year in the electronics field, one of the many outstanding achievements, and possibly the most interesting, was the application of the "solar battery" to the practical production of useful electrical power directly from solar energy. The solar battery was taken out of the experimental classification during the year, when it was installed to supply power to charge telephone batteries on a telephone line near Americus, Georgia.

Another year of study of the transistor has produced many more useful production types of this important device. Production types of many new tubes such as high-power klystrons, traveling-wave and backward-wave tubes, and subminiature tubes became available during the year. Production of the new type high power silicon rectifier was also announced this year.

The results of much basic research in the field of electronics has been revealed. Recent investigations of thermionic and field-emission cathodes promise to revolutionize the vacuum-tube art. Considerable progress in the study of radio propagation has been made, as evidenced by the installation and successful operation of many "forward-scatter" radio links.

Electronic instrumentation in industry and in nuclear science has also made great srides during the year.

As the electronics field has diversified, so have the opportunities in the field diversified for the well-trained technical man. This is evidenced by a greater demand than ever for qualified technical personnel. 1956 and the years ahead hold great promise for those prepared to stay abreast of the developments in the electronics field.

The editorial staff of the BULLETIN extends a most sincere Christmas greeting with wishes for success in the year ahead.

APPLICATION NOTES ON THE PHILCO SURFACE BARRIER TRANSISTOR

by Albert L. Cavalieri, Jr.
Sales Engineer, Government and Industrial Division

(Editor's Note: The author wishes to acknowledge the assistance and guidance given him by James B. Angell and Frank P. Keiper in the preparation of this article.)

THERE IS NO DOUBT that the most significant addition to the field of electronics in recent years is the transistor. Although in its earliest form (the point-contact transistor) it showed great promise in reducing size, weight, and power consumption of many types of electronic equipment, the transistor suffered some inherent limitations such as poor noise figure and instability under certain load conditions. The junction transistor practically eliminated these difficulties, and has been made available in a wide range of power-handling capacities. The cost of these improvements was a rather severe restriction on the upper frequency limit. Recognizing that if transistors were to come of age, the upper frequency limit would have to be considerably extended, and also recognizing the incumbency of low cost, reliability, and reproducibility on their design, Philco set out to develop a new transistor. The result was the Surface Barrier Transistor—a low-power high-frequency transistor. It is formed by electrolytically etching and subsequently plating emitter and collector electrodes upon a suitable base material. In the case of the SB-100 transistor, this material is germanium. By virtue of the process used, a close control on base thickness and electrode size is maintained. As a result, production transistors are remarkably uniform. In addition, the process lends itself well to automation with its ultimate minimum cost.

SB-100 designates a general purpose Philco Surface Barrier Transistor for fast-switching and high-frequency use. The design of this unit was greatly influenced by the undeniable truth that no one transistor will do every job and do it well. In recognition of this fact, the α_{fb}* of the SB-100 has been optimized for general high-frequency use. It is not so close to unity that the grounded emitter cutoff, $f_{\alpha e}$, is drastically low, nor is it so far from unity that useful gain is low. Similar consideration affected C_c which, from gain and bandwidth considerations, should be as low as possible. To reduce C_c, the collector electrode size is reduced and the base thickness increased. Both of these variations tend to reduce α_{fb}. The present design leads to both good gain and bandwidth performance.

The result of this sort of design is indicated by the data sheet value of the average useful upper frequency limit—45 mc — which shows the clear-cut superiority of the Surface Barrier Transistor over junction transistor triodes. Clearly, α_{fb} is not as high as might be desired for some low-frequency applications, but Surface Barrier Transistor performance in high-frequency and switching applications is unchallenged.

The SB-100 is designed for low-voltage applications. It is especially well suited for applications such as computers, where large numbers of high-speed devices are involved, making small size and, above all, low power consumption important. Through its ability to operate at 4.5 volts or less with very little current, the Surface Barrier Transistor effects real power savings.

With regard to the maximum value of collector voltage, a few remarks are

*Definitions of terms are given at the end of this article.

in order. This maximum is called V_{max}, and is defined as the voltage at which the common-emitter open circuit output impedance is one-half its typical operating-point value. There are three causes leading to V_{max} effects. These are: an increase in α_{fb} to values approaching, equalling, or exceeding unity; impurities in or on the germanium; and an increase in the collector barrier width such that the *effective* base thickness reaches zero (referred to as "punch-through," which is a virtual short circuit). With regard to the first of these effects, α's of unity are never encountered in the SB-100. However, as collector voltage increases, α increases slightly from its original value; and, since the output impedance is related to $(1-\alpha_{fb})$, rather significant changes in the output impedance can result. This effect, which is common to all transistors, occasionally influences V_{max} for Surface Barrier Transistors. Of greater importance is the effect of impurities on V_{max} and transistor life. The mechanism by which these impurities affect V_{max} is not clearly understood. However, it is known that transistor life is determined mainly by the surface contamination level and the physical arrangement of contaminants. Roughly, the collector voltage tends to arrange contaminants in an unfavorable way so as to impair performance through a decrease in r_c and an increase in I_{co}. High voltage and temperature increase the speed of this process. The very strictly controlled manufacturing techniques used result in long useful life provided the data sheet limits of V_{max} and temperature are observed. The third determining factor of V_{max}, punch-through, is rarely limiting.

As far as maximum collector current is concerned, it seems that any value obtainable without exceeding some other specifiçation is permissible. The data sheet value for collector current is that value above which α_{fb} drops seriously with consequent increase in drive requirements. Such change in α_{fb} may be determined from the common-emitter collector characteristics.

SURFACE BARRIER TRANSISTOR AMPLIFIERS

The Surface Barrier Transistor is ideally suited to a wide range of applications. Among these are low-level video amplification, high- and medium-frequency amplifiers and oscillators, frequency converters, amplitude detectors, high-speed switches, and possibly high-current pulse generators. The common feature of all these is high-frequency or high-speed performance. Many details, such as the ease of neutralizing and the highest frequency of use, affect the choice of circuit. Therefore, TABLE 1 has been included to show some typical transistor characteristics and performance in what is believed to be the most suitable connection for each application.

The input and output impedances of transistors are functions of frequency, current, and to some extent, voltage. No attempt is made here to present exact theoretical expressions of their nature because of their complexity. The expressions shown in TABLE 2 will hold to a first approximation.

Video Amplifiers Without Feedback

A two-stage Surface Barrier Transistor video amplifier is shown in figure 1. With all the coils shorted, the amplifier is a resistance-capacitance-coupled common-emitter circuit. In the circuit shown, the transistor load impedance is esssentially the same as its input impedance. For this condition the theoretical voltage gain-bandwidth product is $\alpha_{fb} f_{\alpha b}$, because the current gain, and hence the voltage gain, is equal to

$$\frac{\alpha_{fb}}{1-\alpha_{fb}}$$

and the bandwidth is equal to

$$(1-\alpha_{fb})\, f_{\alpha b}.$$

Due to input and output capacitance and power loss in the interstage resistors, the measured value of gain-bandwidth is

TABLE 1

TYPICAL MEASURED SBT CHARACTERISTICS AND PERFORMANCE

INPUT IMPEDANCE	OUTPUT IMPEDANCE	POWER GAIN	BANDWIDTH	OPERATING CONDITIONS	TRANSISTOR CONNECTION
3K, $150 \mu\mu f$	50K, $15 \mu\mu f$	33 db	$\approx$ 10 kc.	$V_c = 3.0$v $I_c = 0.2$ ma Neutralized	Grounded emitter 455-kc r-f amplifier
400Ω, $10 \mu h$	1.5K, $6 \mu\mu f$	6-9db	10 mc.	$V_c = 3.0$v $I_c = 0.5$ ma Neutralized with $r_n = r_b'$; $C_n = C_c$	Grounded base 30-mc. r-f amplifier
1 K, $250 \mu\mu f$	20K, $18 \mu\mu f$	40 db 14 db	2 mc. 12 mc.	$V_c = 3.0$v $I_c = 0.5$ ma Load impedance equal to input impedance	Two-stage grounded-emitter iterated video amplifier

TABLE 2

APPROXIMATE INPUT-OUTPUT IMPEDANCE

INPUT IMPEDANCE	OUTPUT IMPEDANCE	OPERATING CONDITIONS	TRANSISTOR CONNECTION
$* r_d + (2-a) r_b'$	$\frac{r_b' - jx_{c_c}}{2}$	Dissipative neutralization (See figure 3.)	Common base
$r_b' + \frac{r_d}{1-a}$	$\frac{1}{\frac{1}{R_C(1-a)} + j\omega \left(\frac{c_c}{1-a}\right)}$	Lossless neutralization $f \ll f_{ab}$	Common emitter

$* r_d = \frac{kT}{Q i_e}$ $\qquad r_d = \frac{27}{i_e}$ at room temperature with i_e in milliamperes

K = Boltzman's constant
T = temperature (absolute)
Q = electron charge
i_e = emitter current

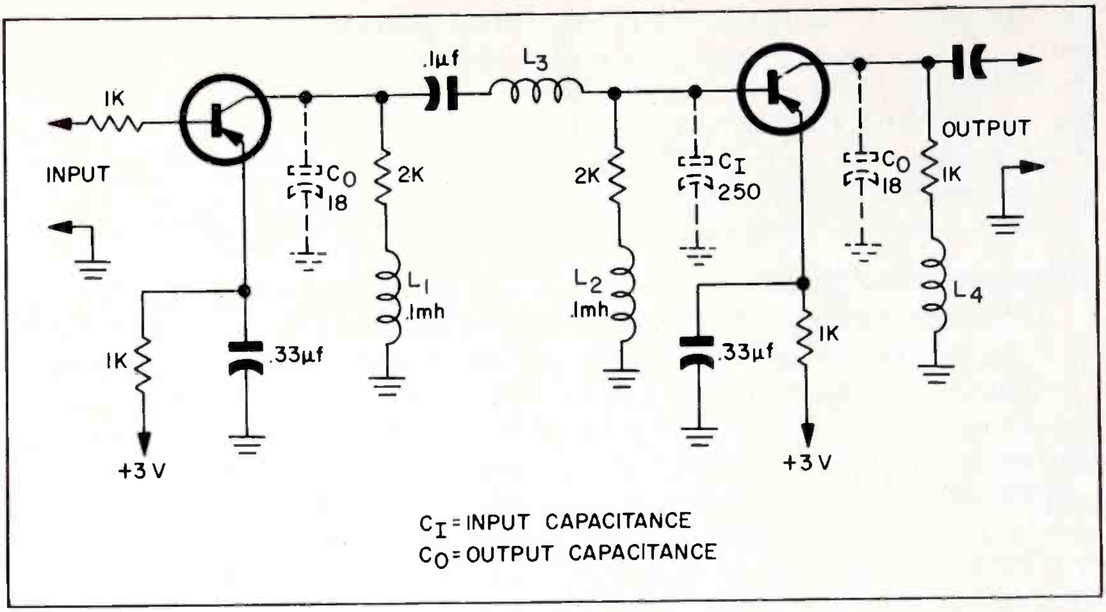

Figure 1. Video Amplifier Without Feedback

lower than the calculated value. For a transistor with an $f_{\alpha b}$ of 50 mc and a low-frequency α_{fb} of 0.90, a stage gain of 5 and a bandwidth of 5 mc. are considered typical. The addition of L_3 and L_4, chosen to resonate with the ouput capacitance at the upper frequency of interest, will improve the gain bandwidth by 50 percent. L_1 and L_2 improve the product by another 30 percent by decreasing the loading effect of the resistors at the high end. These values are so chosen as to affect the gain at 3 mc. and above. Thus, with proper coils, a gain bandwidth nearly equal to $f_{\alpha b}$ is achieved.

Video Amplifiers With Feedback

The gain-bandwidth can also be made to approach the theoretical limit of the Surface Barrier Transistor through the use of a series peaking coil and feedback. The bandwidth may be made to assume a wide range of values by choice of the feedback resistors R_{f1} and R_{f2}, shown in figure 2. With transistors having an alpha cutoff of 45 mc., the two-stage

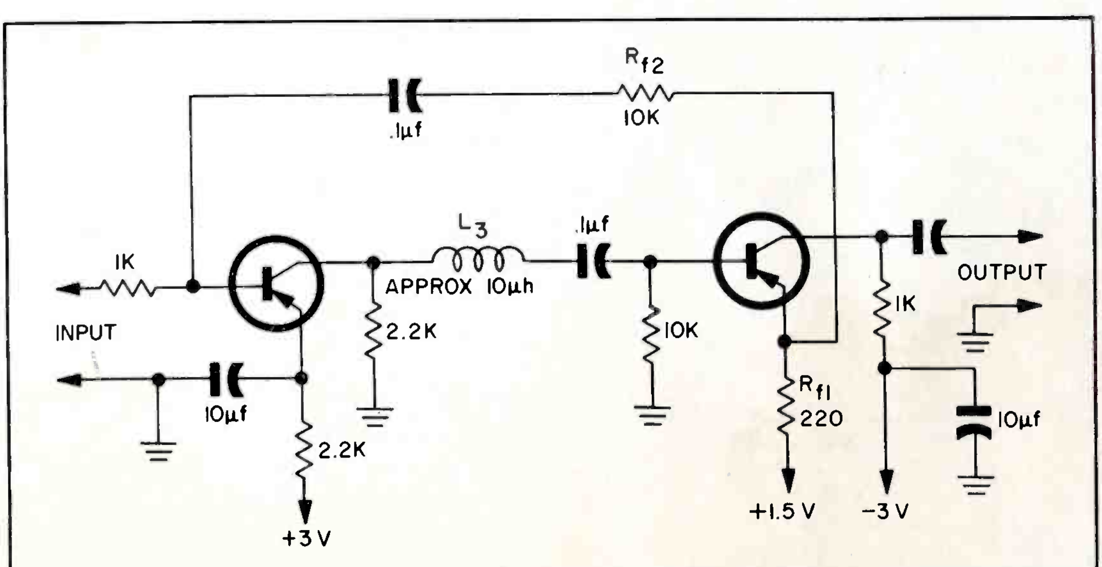

Figure 2. Video Amplifier Employing Negative Feedback

bandwidth is variable from 2 to 12 mc. with a corresponding gain variation from 40 db to 14 db. Cascading of such iterated feedback amplifiers is possible without interaction because the feedback is independent of load provided true iteration is maintained in the cascade.

Bandpass Amplifiers

At frequencies higher than about 0.1 $f_{\alpha b'}$, the usual transistor amplifier connection is the grounded-base circuit shown in figure 3. With this type connection and with proper neutralization, amplification at frequencies close to f_{max} can be achieved. f_{max} is the highest frequency at which the transistor can be made to oscillate. The minimum power gain to be expected is:

$$G_p = \frac{(f_{max})^2}{f} \qquad (1)$$

where

$$f_{max} = \sqrt{\frac{\alpha_{fb} f_{\alpha b}}{8\pi r_b' C_c}} \qquad (2)$$

It is desirable, though not absolutely necessary, to neutralize such stages to prevent oscillations and to simplify tuning. The components R_n and C_n effect complete neutralizing by forming a bridge circuit with r_b' and C_c. This method of neutralizing is dissipative and the loss must be included in gain calculations. However, only near f_{max}

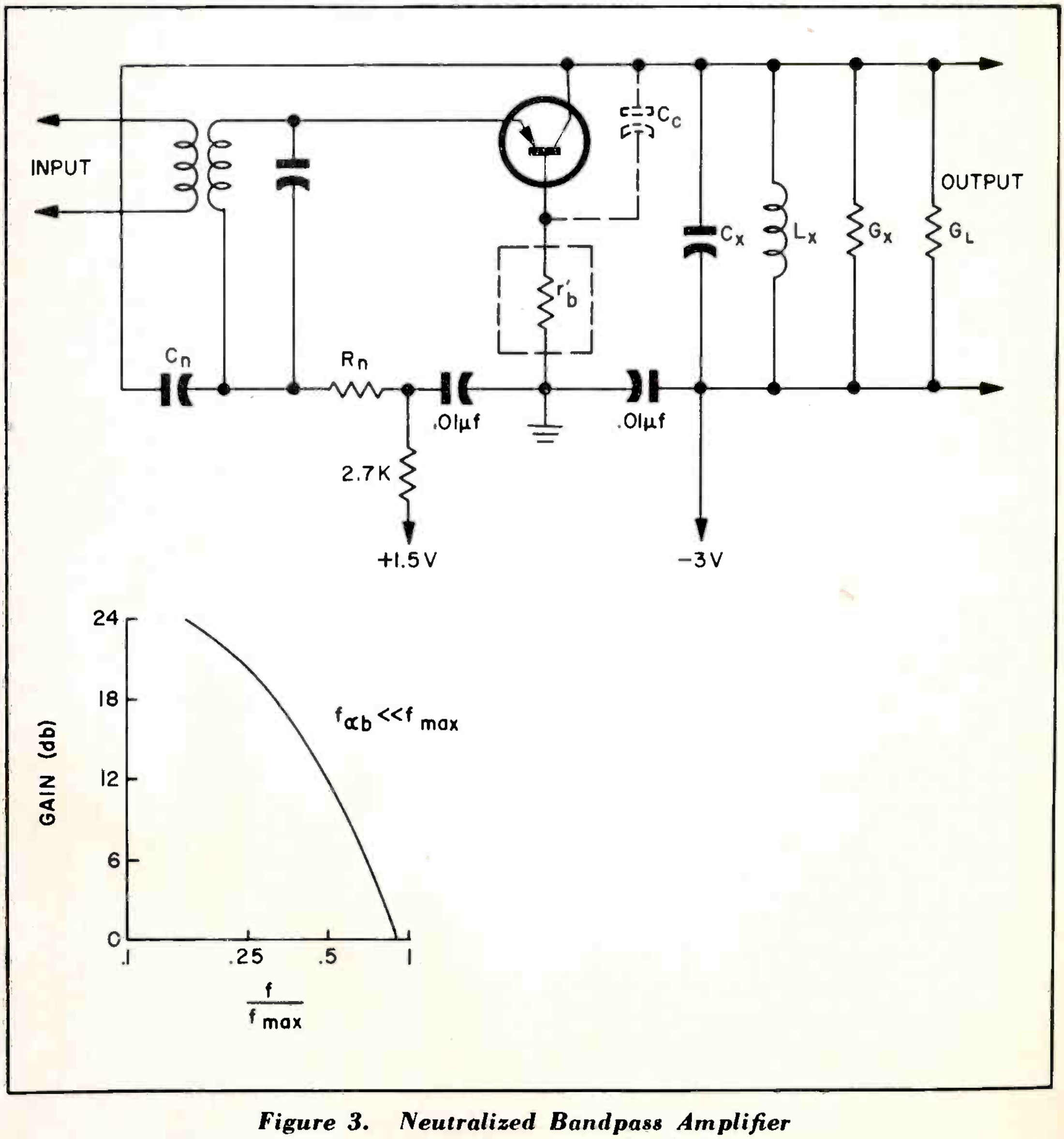

Figure 3. Neutralized Bandpass Amplifier

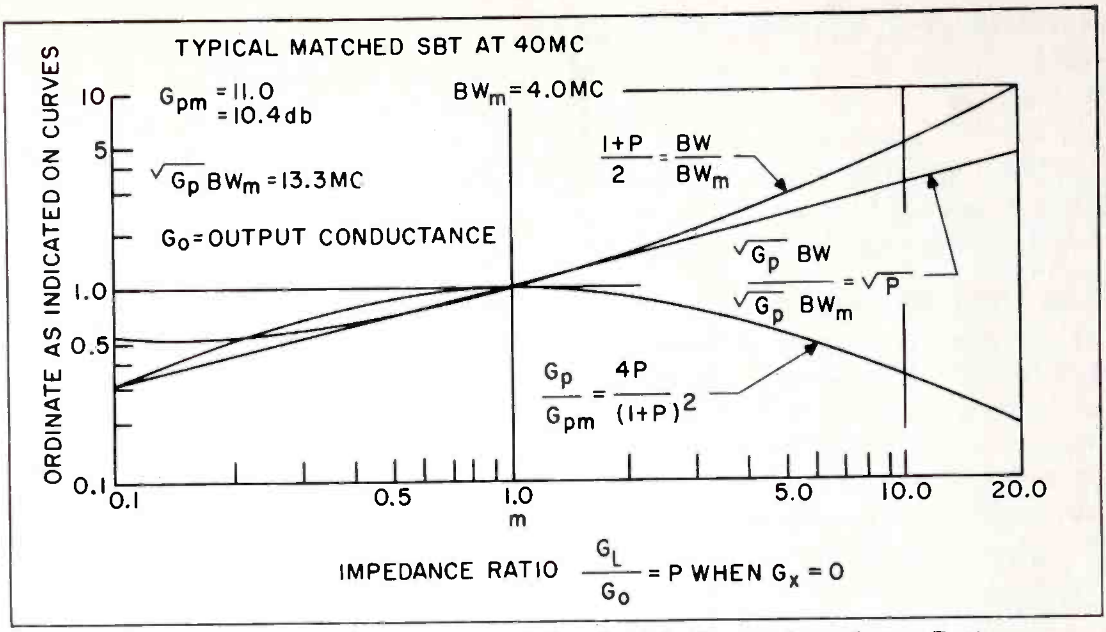

Figure 4. Normalized Gain and Bandwidth Versus Impedance Ratio

does the loss reduce the realizable gain to even slightly below that expressed by equation (1). The conditions $r_b = R_n$, $C_c = C_n$ result in the smallest power gain loss for dissipative neutralizing. The bandwidth of such a stage is:

$$BW = \frac{G_o + G_x + G_l}{2\,C_c + C_x} \tag{3}$$

where G_o = output conductance and the power gain is:

$$G_p = \frac{\alpha_{fb}^{\,2}}{8\,(2\pi f C_c)^2\, r_b'\, r_{in}} \tag{4}$$

where r_{in} = resistive component of input impedance.

Under matched conditions and with no external capacitance added, the voltage gain-bandwidth product referred to the same impedance levels is:

$$A{\cdot}BW = \frac{\alpha_{fb}\, f_o}{2.8} \tag{5}$$

where f_o is the center frequency.

As indicated in figure 4 substantial improvements in gain-bandwidth product can be effected through mismatching the load. Nevertheless, equation (5) provides a convenient figure of merit for high-frequency wideband tuned amplifiers.

Somewhat higher gain can be obtained at frequencies below about 0.1 f_{max} by using the grounded-emitter connection shown in figure 5. The neutralizing capacitance, C_n, is made equal to C_c times the turns ratio of the transformer in the collector circuit. Thus it neutral-

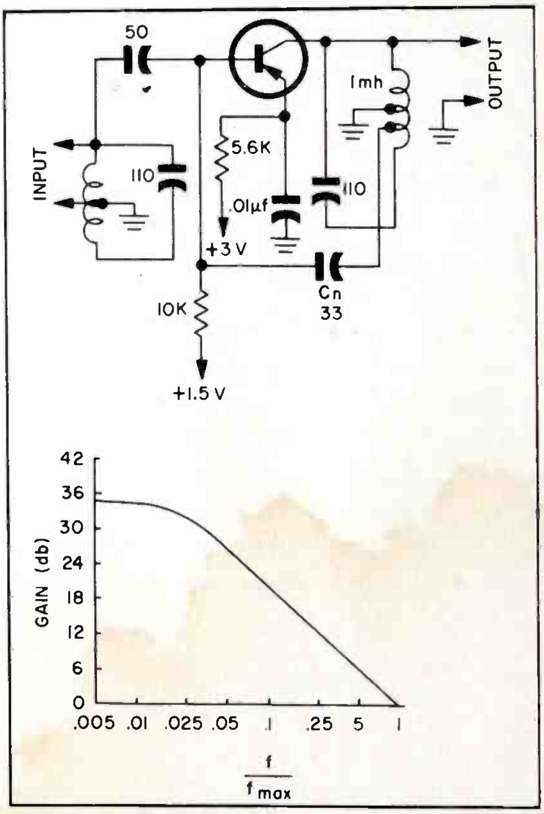

Figure 5. Neutralized Common Emitter Tuned Amplifier at 455 Kc.

izes the effect of C_c. Such partially neutralized amplifiers yield gains of about 33 db at 455 kc. At still lower frequencies, below about 200 kc., the low-frequency parameters take over, giving gains of about 35 db.

Experience has indicated that it is important to keep the coupling in all coils high, especially where the coils have one or more taps. Unless this condition is met, it is possible for spurious responses or oscillations to occur. Even the use of ferrite rods has sometimes been inadequate. However, if ferrite rods are used, and the winding length is kept short, no problems should be encountered.

The grounded-emitter gain characteristic in the matched case

$$\left(\text{where } r_1 = r_c \left[1 - \alpha_{fb}\right]\right)$$

is nearly independent of the α_{fb} of the transistor used. The expression for approximate gain is:

$$G_p \approx \frac{r_c}{4r_d}. \qquad (6)$$

This expression of gain is approximately correct even for the widest variations of α that may be encountered in Surface Barrier Transistors.

HIGH FREQUENCY OSCILLATORS

Since an oscillator is nothing more than an amplifier with appropriate feedback to sustain oscillations, the Surface Barrier Transistor is also well suited to high-frequency oscillator service. Regardless of the way in which the transistor is connected, f_{max} is given by equation (2). As a result of a large number of measurements, the efficiency of oscillation of the Surface Barrier Transistor at a frequency f has been found to be:

$$N = 0.8 \log_{10} \frac{f_{max}}{f} \qquad (7)$$

This efficiency is defined as the ratio of the useful oscillator output to the power supplied from the collector supply. An oscillator circuit is shown in figure 6.

HIGH-SPEED SWITCHING

The excellent high-frequency performance of Surface Barrier Transistors makes them well suited for high-speed switching. A basic circuit consists of a collector-coupled common emitter flip-flop, which is shown in figure 7. The trigger requirement is a positive pulse of one-half to one volt. The steering diodes, D_3 and D_4, guide the triggering pulse to the proper transistor. When TR-1 is conducting, the cathode of D_4 is at a positive potential and is cut off. Since D_3 is biased in the forward direction, it passes the input pulse and TR-1 is cut off. This action turns on TR-2. The next trigger pulse cuts off TR-2 and causes TR-1 to conduct. The diodes D_1 and D_2 prevent the collector from reaching too low a voltage during conduction. In this way they prevent accumulation of a large number of holes in the base region. These holes would have to be collected during the off time and thus would adversely affect the pulse fall time. The output of the circuit as shown

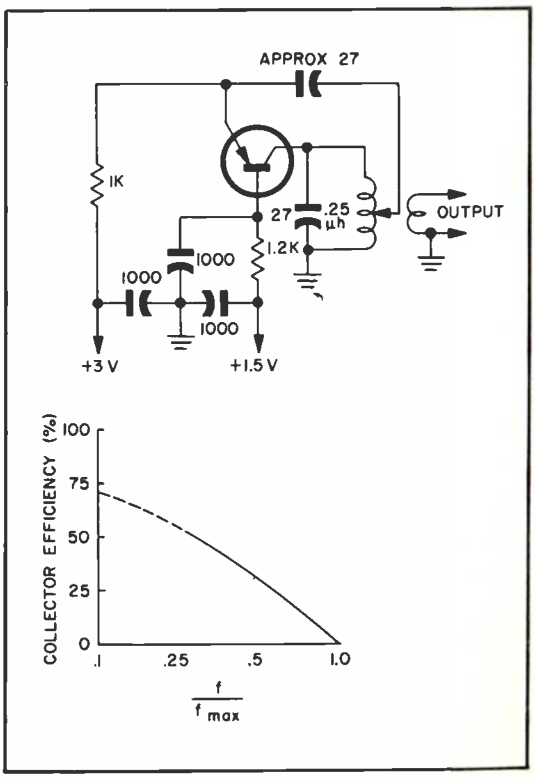

Figure 6. High-Frequency Oscillator (50 Mc.)

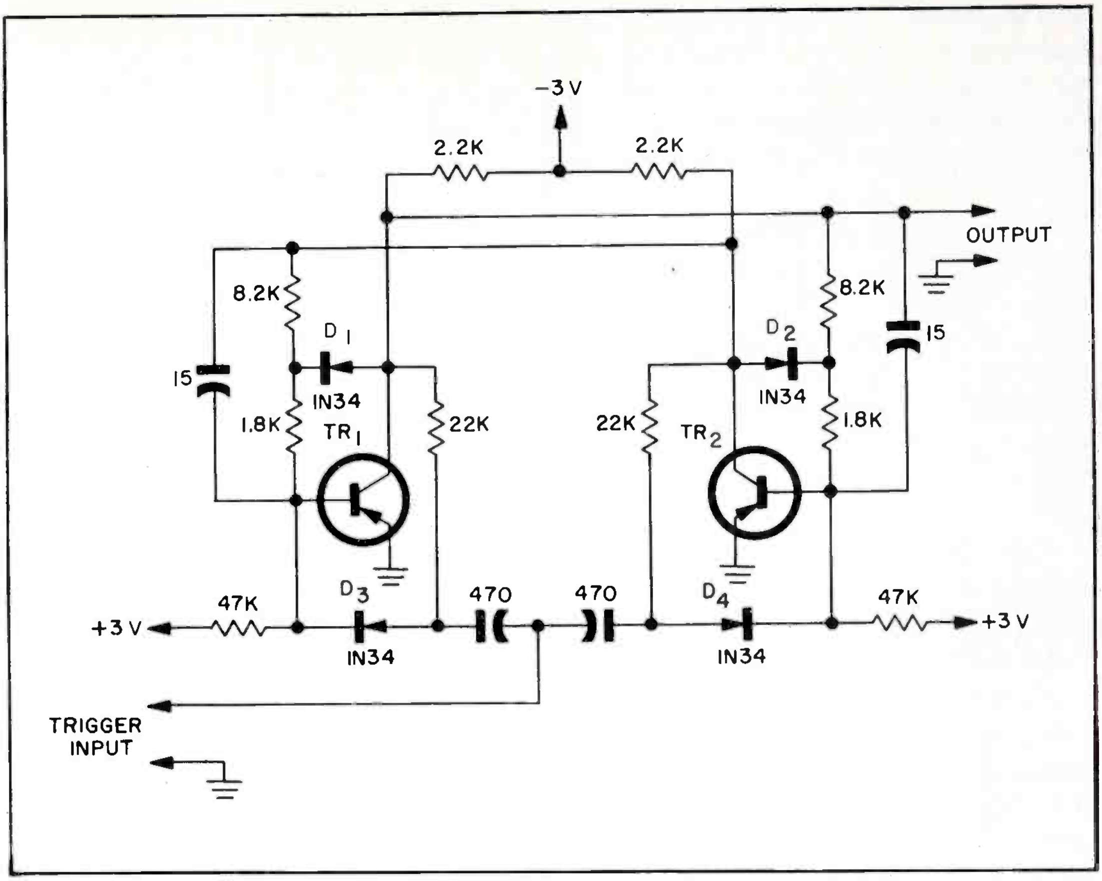

Figure 7. Schematic of Bi-stable Flip-Flop

is a 2- to 3-volt pulse with a rise and fall time of less than one-tenth microsecond.

PULSE GENERATION

The circuit in figure 8 has been used for low-duty-cycle, short-rise-time pulse generation. As an answer to a laboratory requirement using parts on hand, it performed well enough to be worthy of note. At a nominal 2-kc. repetition rate, the circuit produced pulses of 3 microseconds duration with 0.6 watt peak power. The rise and fall times of the pulses were less than 0.2 microsecond. After several hundred hours of use, no detectable change in operation was observed.

The use of voltages as high as those shown is not recommended in general. The transistor used was selected to have a "punch-through" voltage of 10 volts, which enabled it to operate in this circuit. However, excellent operation at lower potentials can be obtained through the use of a more suitable transformer (one designed with lower inductance in the primary and secondary windings).

The peak emitter current, on the order of 100 ma. in the application shown, does not lead to deterioration provided other limits are observed. Emitter currents in excess of 130 ma. have often

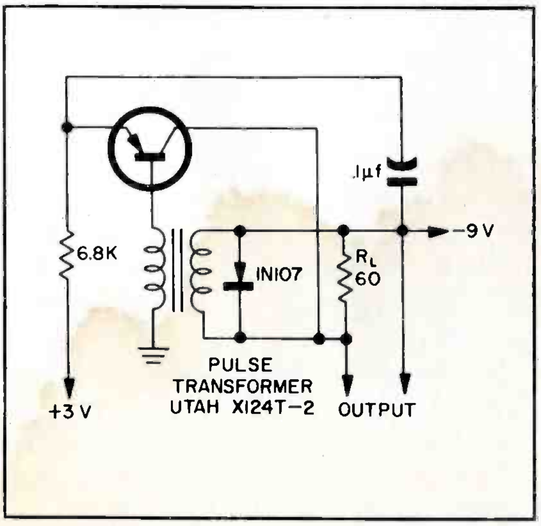

Figure 8. Pulse Generator

been measured in circuits; it seems that if the transistor is kept within rated dissipation and maximum voltage limits, and is driven hard enough, much greater peak currents can be safely passed.

CONVERTERS

Converter action in the broadcast frequency range is readily obtained by means of the circuit shown in figure 9. In this circuit, the input and local-oscillator signals are mixed in the emitter diode. The collector is tuned to the difference frequency resulting from this mixing, and is thus grounded at local-oscillator and signal frequencies. Grounded-collector current amplification occurs between base and emitter at these two frequencies. At the difference frequency the emitter and base are essentially grounded so that a gain from emitter to collector occurs. The conversion gain of a typical stage is 20 db with a noise figure of 4 db.

The feature of the SB-100 that suits it for broadcast-band converter service, and indeed for amplifier service too, is its utility in a common-emitter or common-collector connection in the frequency range up to 3 megacycles. Connected in the same manner and operated over the same frequency range, a junction transistor will usually exhibit only low gain, for although the low-frequency value of alpha is high, the alpha cutoff frequency is relatively low. Since the common-emitter cutoff frequency is expressed as $f_{\alpha b}(1 - \alpha_{fb})$, it is obvious that if gain is required at high frequencies, $f_{\alpha b}$ must be high. This condition is met in the Surface Barrier Transistor. In addition to the above, drive requirements vary with α_{fb}. As a result of the high cutoff frequency, the drive requirements of the Surface Barrier Transistor are maintained low and substantially constant.

AMPLITUDE DETECTORS

A process very similar to that described above occurs in the amplitude detector. In this circuit, figure 10, a gain of 18 db can be realized in the detection of frequencies up to a few megacycles, such gain being defined as the ratio of audio to sideband power. The remarks made relative to the effect of $f_{\alpha b}$ on converter performance apply directly to amplitude detectors.

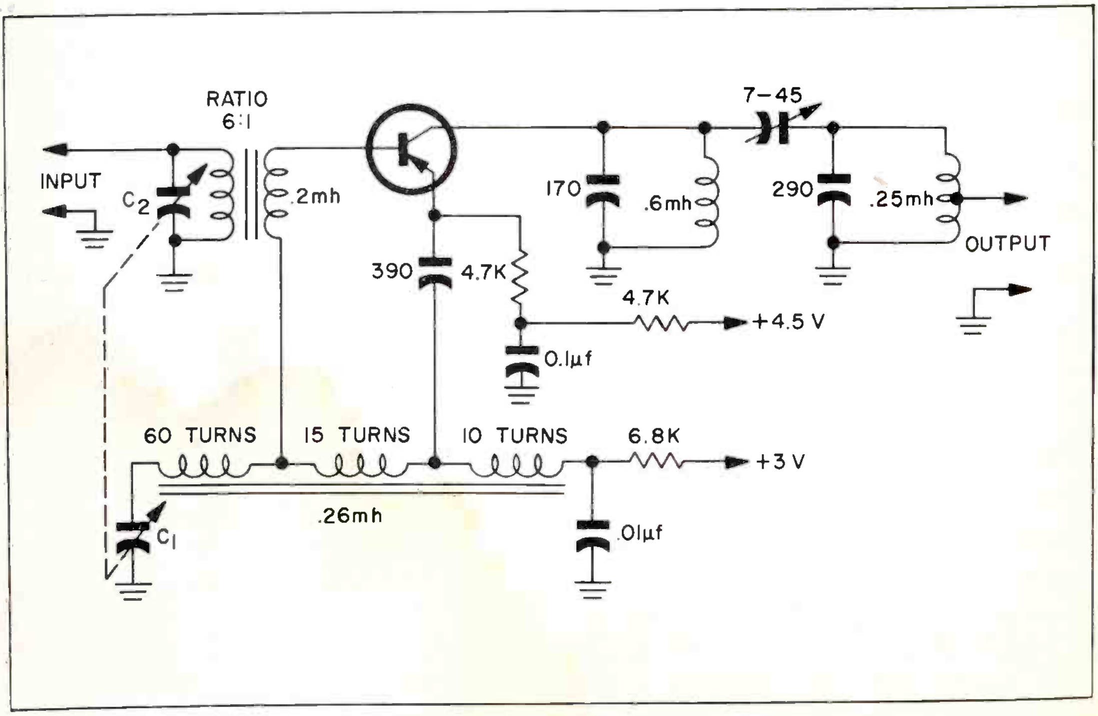

Figure 9. Broadcast Band to 455 Kc. Converter

PRECAUTIONS

Although the SB-100 is rugged, it can be damaged by abusive treatment. Such treatment may accidentally occur through the connection of the transistor with test equipment having an input lead which is at an elevated potential due to a charged blocking capacitor. Or, the equipment may be such as to allow the passage of large leakage currents from the a-c line to the circuit. Notable in this class is the soldering iron. It is recommended practice to use gun-type irons or isolation transformers with conventional irons. In addition, attention should be given to proper grounding. Such precautions properly observed will ensure long, satisfactory transistor life.

DEFINITIONS OF TERMS

α = Gain.

α_{fb} = Common base forward current gain.

C_c = Collector capacitance.

$f_{\alpha e}$ = Frequency at which common emitter current gain is down 3 db.

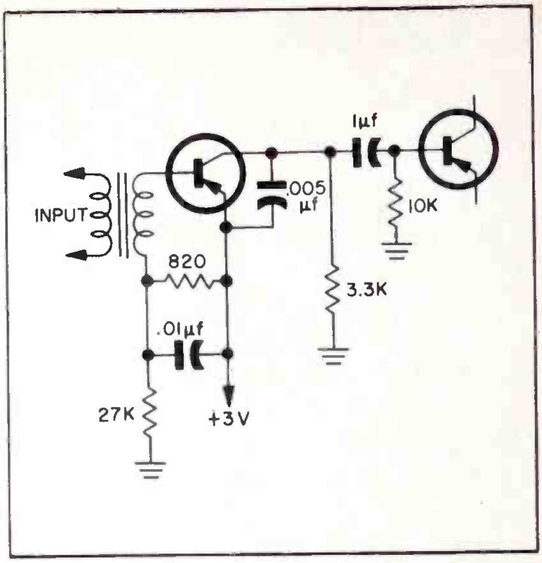

Figure 10. Amplitude Detector

$f_{\alpha b}$ = Frequency at which common base current gain is down 3 db.

r_c = Collector resistance.

I_{co} = Collector cutoff current.

G_{pm} = Power gain for matched condition.

BW_m = Bandwidth for matched condition.

Solution to . . . September-October "What's Your Answer?"

Although the circuit of V_2 in figure 1 resembles a cathode follower, it is actually a cathode-loaded amplifier. Since the input to V_2 is developed across the secondary of T_1, which is directly between the grid and cathode of V_2, the plate current of V_2 flowing through the cathode resistor causes no degeneration. Consequently, the voltage amplification of V_2 is exactly the same as if the 20K resistor were placed in the tube's plate return to B+, as in a conventional grounded-cathode amplifier.

In figure 2, on the other hand, the input to V_2 is developed across the transformer winding of T_1 which is used as the d-c plate return for V_1. Since R and the cathode resistor of V_2 are in series across the transformer winding, it is apparent that the voltage across the cathode resistor (the output voltage) can never exceed the voltage across the transformer winding (the input voltage). The circuit of V_2 in figure 2 is actually a cathode follower and develops a voltage amplification which is less than 1.

RADAR RECEIVER A-F-C SYSTEMS
PART 3

by H. W. Merrihew
Headquarters Technical Staff

This is the third article of a series of four on radar a-f-c systems. The first two articles appeared in the two previous issues of the BULLETIN. For a thorough understanding of the following material, the reader should be familiar with the material contained in the previous articles.

AFC FOR A THERMALLY TUNED KLYSTRON

The a-f-c circuit, shown in figure 1, controls a thermally tuned 2K45 klystron. This tube is used as an oscillator at X-band frequencies and actually consists of two tubes, the klystron and a triode referred to as a *heat triode.* The plate temperature of the heat triode determines the size of the klystron cavity. When the grid of the heat-triode section is at zero bias, its plate heating is greatest and the klystron cavity is expanded to lower the frequency. In the circuit of figure 1, maximum heating occurs when the right-hand section of the heat multivibrator is cut off and the grid of the heat triode is held at zero volts by the divider. The heat multivibrator is an Eccles-Jordan circuit. The timing multivibrator of figure 1 is free-running at about ¼ cycle per second and controls the heat multivibrator operation.

A-F-C Search Action

The a-f-c search action, which is controlled by the timing multivibrator, will now be considered. Since minute circuit differences determine which section of the heat or timing multivibrator tubes will conduct the largest current when power is initially supplied, any one of a number of conditions may initially exist; however, the end result is the

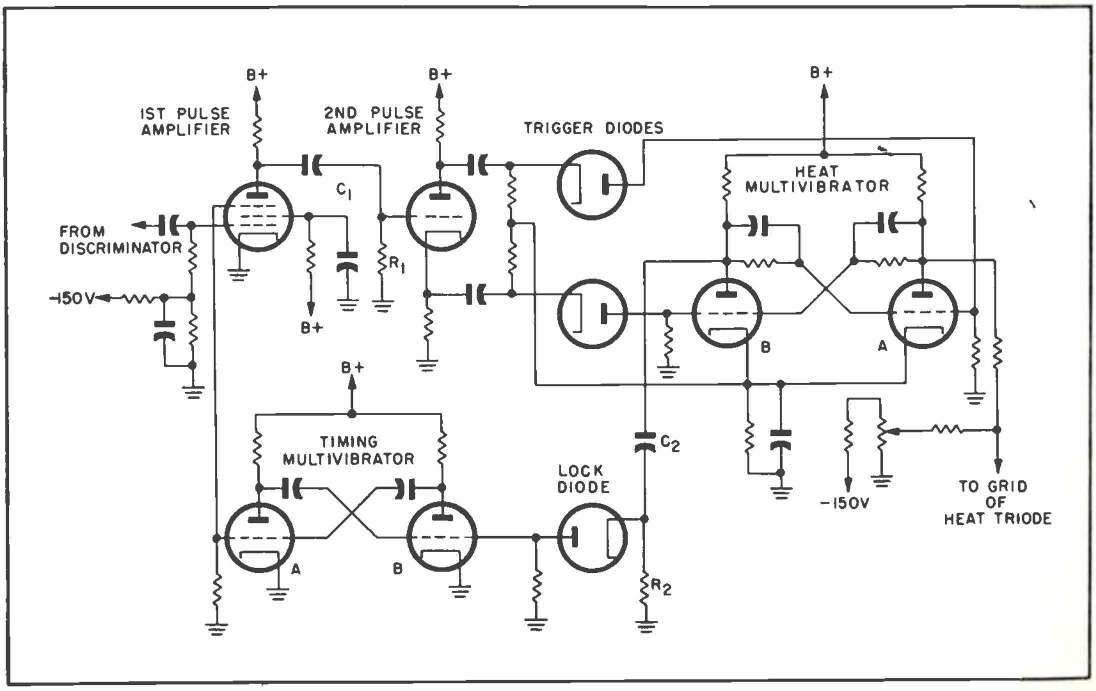

Figure 1. A-F-C Circuit for Thermally Tuned Klystron

same. For discussion purposes, assume that the A section (refer to figure 1) of both the heat and timing multivibrator tubes initially conduct the largest current, and therefore are the "on" tubes. (An interesting feature of the circuit is the way in which the heat and timing multivibrators synchronize, as shown by the synchrogram in figure 2.)

Since the A section of the heat multivibrator is on, the plate is at its maximum negative excursion; and since the A section of the timing multivibrator is on, the grid is at its maximum positive excursion. The A section control grid of the timing multivibrator is connected to the suppressor grid of the first pulse amplifier, causing it to draw plate current. The plate signal is shown as the fifth waveform of the synchrogram in figure 2. This negative-going wave is differentiated by C_1 and R_1 and appears at the grid of the second pulse amplifier as a negative pulse (sixth waveform of figure 2). This pulse is applied through the lower trigger diode to the grid of the B section of the heat multivibrator. Since the B section is already off, the pulse has no effect, and no further action occurs until the timing multivibrator "turns over." When this happens, a positive-going wave appears at the plate of the first pulse amplifier because the suppressor grid is driven negative as the A section of the timing multivibrator is driven to cutoff. This wave is differentiated by the combination of R_1 and C_1 and applied through the second pulse amplifier and trigger diode to the grid of the A section of the heat multivibrator, to turn it off and to turn the B section on.

For the condition discussed above, the heat triode of the klystron is initially cut off, and when the timing multivibrator "turns over," the grid of the heat triode changes from cutoff to zero bias, and the triode heats the klystron cavity. This process sweeps the klystron frequency from its highest to its lowest value. Note that during this time "lock-in" cannot occur because the suppressor grid of the first pulse amplifier holds this tube at cutoff, and pulses from the discriminator cannot be passed to the heat multivibrator.

The synchrogram in figure 3 shows the important waveforms involved in the search action. During the heating process the klystron frequency drops, but the system cannot lock in because the pulses from the discriminator, which are applied to the first pulse amplifier, will not be passed. During the heat "off" period, the klystron frequency rises, and

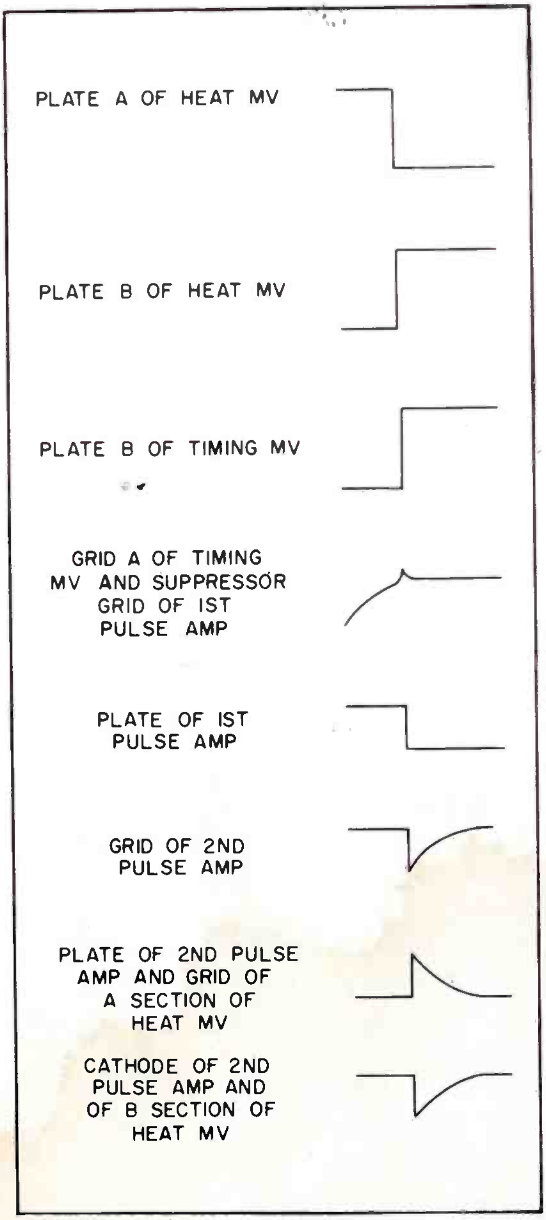

Figure 2. Waveforms for the A-F-C Circuit of Figure 1

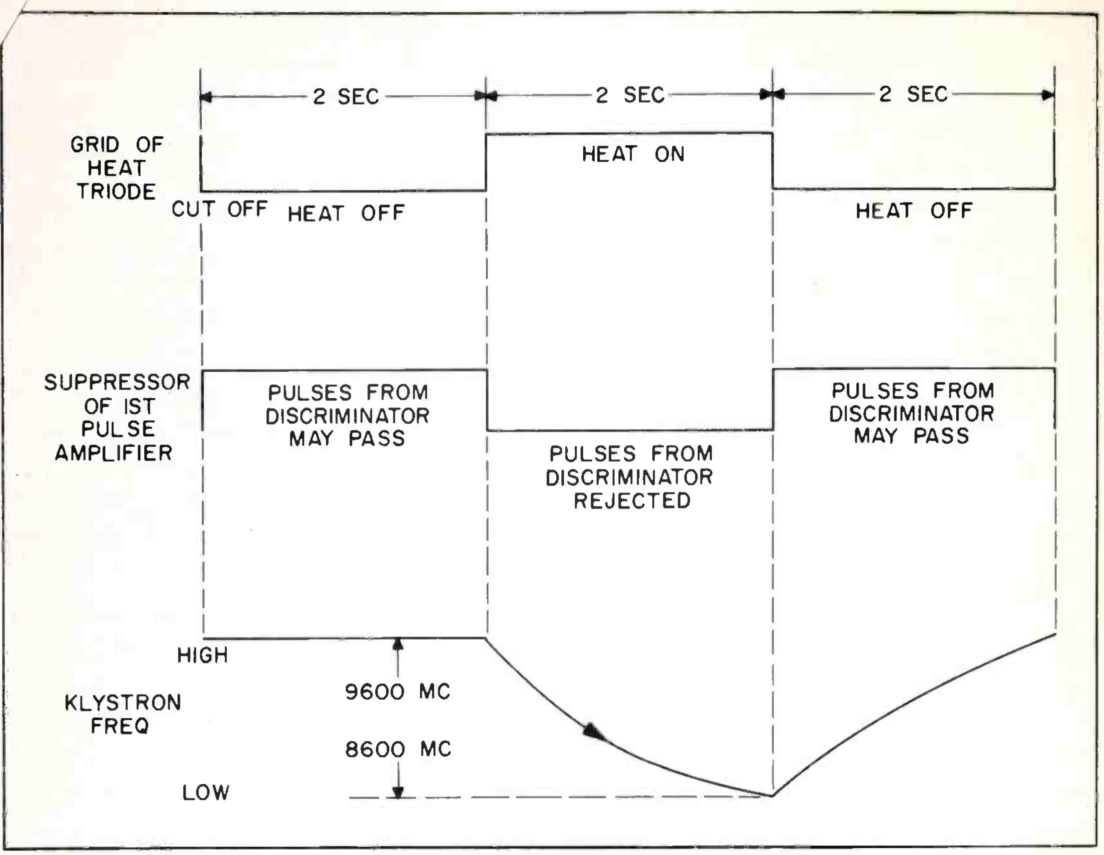

Figure 3. Variation of Klystron Frequency with Heating

lock-in may take place as the klystron output sweeps from a low to a higher frequency.

AFC Locking In

The remaining feature to be examined is the locking-in process. Since the A section of the heat multivibrator is on, the B section must be turned on in order to stop the increase in klystron frequency. As the klystron frequency is raised, the first pulses arriving from the discriminator are positive. The first pulse amplifier inverts these pulses and applies them to the second pulse amplifier. The cathode output of the second pulse amplifier is coupled by the trigger diode to the grid of the B section of the heat multivibrator. Since the B section is already cut off, this trigger has no effect. As the i-f signal frequency passes through 30 mc., the output of the discriminator reverses polarity, and negative pulses are applied to the first pulse amplifier. The resulting negative output at the plate of the second pulse amplifier is coupled by the upper trigger diode to the grid of the A section of the heat multivibrator. This cuts off the A section, which turns on the heat triode of the klystron. As soon as the klystron frequency drops enough to allow the i-f signal to pass through 30 mc., positive pulses again appear at the grid of the first pulse amplifier. The cathode output of the second pulse amplifier is again passed by the lower trigger diode to cut off the B section of the heat multivibrator. This rapid reversing (between 100 and 200 times per second) keeps the plate temperature of the heat triode fairly constant, and the desired frequency for the klystron is obtained. Lock-in cannot occur at the high-frequency side of the transmitter signal because the polarity of the output signal from the discriminator is incorrect.

Note that the coupling circuit consisting of R_2 and C_2 supplies a square wave of 50 to 100 pulses per second to the cathode of the lock diode. The negative portions of the square wave are applied to the B section of the timing multivibrator, keeping the B section off, the A section on, and the first pulse amplifier free to pass tracking data to the heat multivibrator. These negative pulses keep enough bias on the timing multivibrator to prevent it from operating. Hence the timing multivibrator is disabled when lock-in occurs.

AFC FOR A LIGHTHOUSE-TUBE OSCILLATOR

The final a-f-c circuit to be discussed, is a motor-tuned type for a lighthouse-tube oscillator. This circuit produces mechanical tuning of the lighthouse tube—no electronic tuning system is involved. The plate and cathode circuits of the lighthouse-tube oscillator are varied simultaneously; the plate circuit functions primarily as a frequency control and the cathode circuit functions primarily as a power output control.

General Circuit Operation

A schematic diagram of the circuit is shown in figure 4. The discriminator is a Travis circuit using capacitive coupling. The discriminator input circuit consists of three separate tuned circuits, one tuned to 29 mc., one tuned to 30 mc., and one tuned to 31 mc. The outputs are balanced at 30 mc., by means of the variable capacitors.

The discriminator output pulses are applied to the pulse stretcher tubes, where they are stretched by the R-C networks in the cathode circuits. The balance control in the grid circuits makes the outputs of the pulse stretchers equal when there is no signal input to the discriminator. The voltage at point A with respect to point B is zero when the input frequency is 30 mc., and any deviation produced by the discriminator will cause point A to become either positive or negative with respect to point B.

The outputs of the pulse stretchers are coupled to the grids of the balanced modulator by lead networks. The lead network in the upper grid circuit is composed of R_1, R_2, and C_1. The lead network in the lower grid circuit is composed of R_3, R_4, and C_2. The lead networks produce excitation signals, at the grids of the balanced modulator tubes, which are proportional to the rate of change of the input signal, as well as to the amplitude of the input signal.

An a-c reference voltage is applied to the plates of the balanced modulator. When the grid voltages are equal (30-mc. signal or no signal at all applied to the discriminator), both tubes in the modulator conduct equally. The modulator is balanced by adjusting the variable resistor in the cathode circuit of one tube.

For discussion purposes, first assume that equal bias voltages appear at the grids of the balanced modulator from the pulse-stretcher circuit. Reference to part A of figure 5 will show that both tubes conduct equally, and that the output of the balanced-modulator transformer is zero. Since the tubes conduct equally, and the currents through the transformer primary flow in opposite directions, the net primary flux is zero. When an i-f signal of less than 30 mc. is applied to the discriminator, balanced modulator tube V_1 conducts more current than V_2, as shown in part B of figure 5, and the net primary flux is such that the resultant a-c output voltage leads the a-c reference voltage applied to the plates by about 70 degrees. The combination of the capacitance and the primary inductance forms a parallel resonant circuit whose resonant frequency is near the a-c reference frequency. The circuit, however, operates at a point far enough from resonance to produce the leading phase angle mentioned above. This leading angle is produced by the capacitor connected across the transformer primary. This capacitor also serves to complete the a-c cycle by the "flywheel" effect it produces. Note

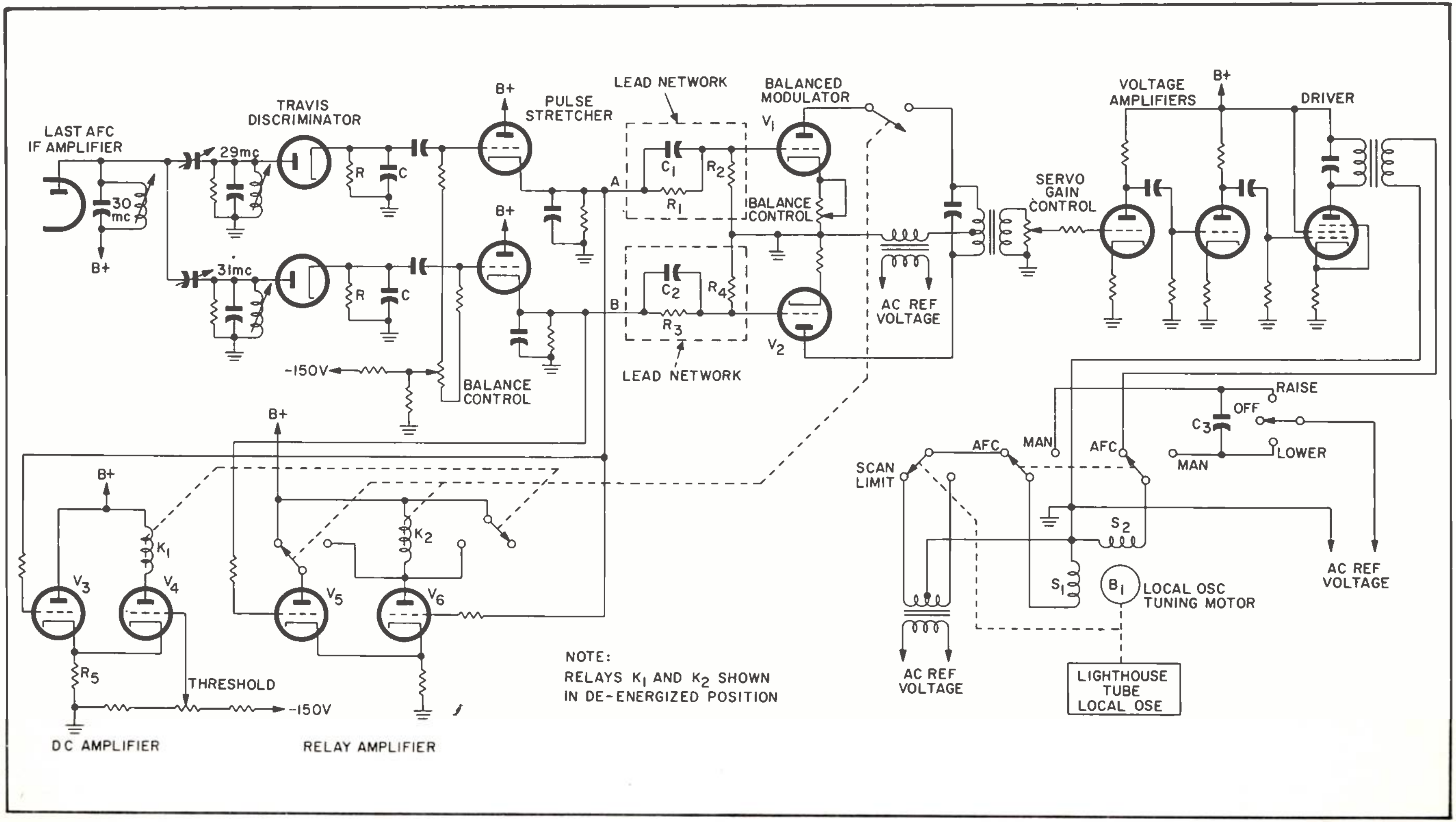

Figure 4. A-F-C Circuit for a Lighthouse-Tube Local Oscillator

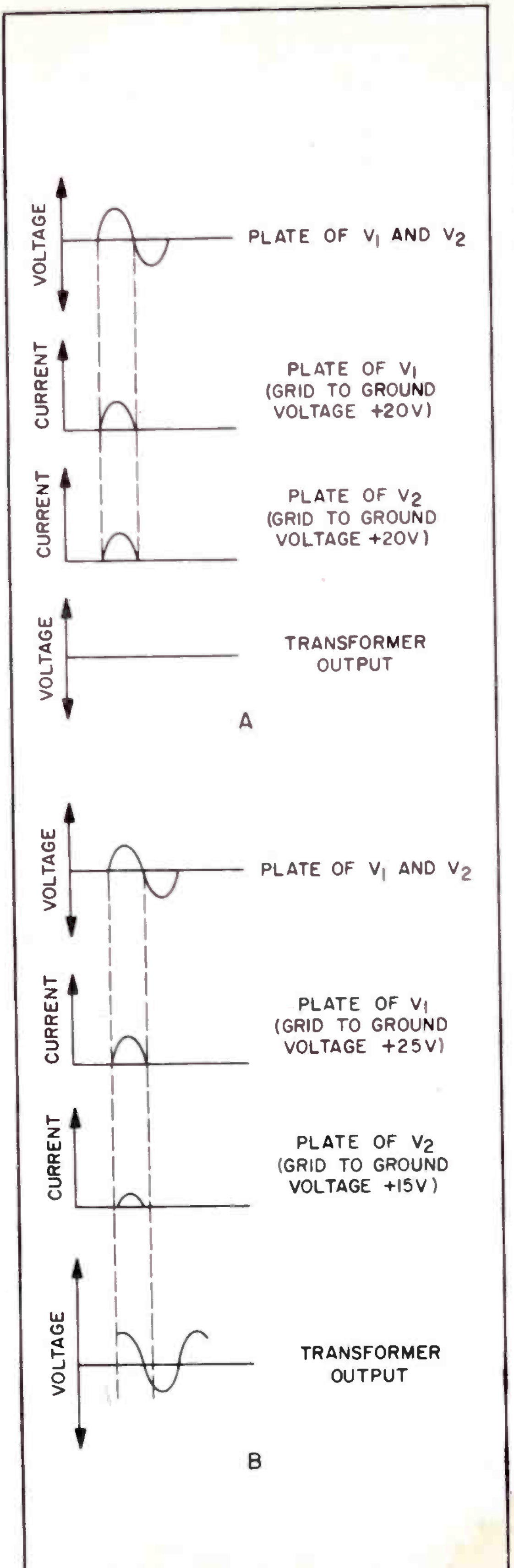

Figure 5. Balanced Modulator Waveforms

that the output would consist of pulses if the capacitor were not present.

When an i-f signal greater than 30 mc. is applied to the discriminator, V_2 conducts more current than V_1, and the net primary flux produces an output signal which is 180 degrees out of phase with the output shown in part B of figure 5.

The balanced-modulator output signal is amplified by the voltage amplifiers and applied to the driver stage shown in figure 4. The driver output supplies one phase of the two-phase winding of motor B_1. The plate circuit of the driver introduces a 20-degree phase shift so that the voltage supplied to the S_2 winding of B_1 is 90 degrees out of phase with the a-c reference voltage applied to the S_1 winding. (Note that the a-c reference voltage applied to the motor is of the same phase as the a-c reference voltage applied to the plates of the balanced modulator tubes.) Hence, an i-f signal of less than 30 mc. causes the motor to rotate in one direction, and an i-f signal greater than 30 mc. causes the motor to rotate in the opposite direction. Manual control of motor rotation is obtained by setting the AFC-MAN switch to the manual position. This inserts capacitor C_3 in series with either winding S_1 or S_2 depending on the position of the RAISE-LOWER switch. The direction of rotation of the motor determines whether the oscillator frequency is raised or lowered.

The Lead Network

When an i-f signal that is different from 30 mc. appears at the discriminator, the output of the pulse stretchers will change. The rate of this change depends upon the drift rate of the intermediate frequency. When the i.f. is drifting rapidly it is necessary to start the motor quickly to follow this change. If the drift rate is slow, the motor may start turning after a greater time.

The lead network is a differentiator circuit which produces a signal whose amplitude is primarily proportional to the rate of change of the pulse-stretcher output signals. An equivalent circuit of the lead network is shown in part A of figure 6, where any motion of the potentiometer arm simulates a changing input voltage. After the initial charge of the capacitor, i.e., after the capacitor charging current has reached a steady value, the lead network produces an output signal which changes in step with the change of input voltage. To illustrate the action of the lead network, a slowly changing pulse-stretcher output and the corresponding lead-network output are shown in part B of figure 6. Part C of figure 6 shows the conditions for a rapidly changing pulse-stretcher output signal.

The magnitude of lead-network signal causes the balanced modulator to have a large or small output, depending mostly upon the rate of change of input signal. Thus, a large signal applied to the balanced-modulator grids provides the motor with a large torque so that it can start rapidly and follow a rapid drift in the i-f signal. For a decreasing input the lead-network output swings downward, as shown in part D of figure 6. In this way the motor is stopped more rapidly and hunting is eliminated.

Search Action

The auxiliary circuits shown in figure 4, consisting of the d-c amplifier, relay amplifier, and associated relays, provide the search feature.

When the system is initially turned on, and no i-f signal appears, the output of the pulse stretchers is zero. This condition permits sufficient current to flow in d-c amplifier V_4 to energize relay K_1, which then de-energizes relay K_2 by placing a short circuit across it, thereby opening the K_2 contact in the plate circuit of balanced-modulator tube V_1. The balanced modulator is thus completely unbalanced, and an a-c voltage is applied to the tuning motor through the voltage-amplifier and driver stages. This voltage drives the tuning

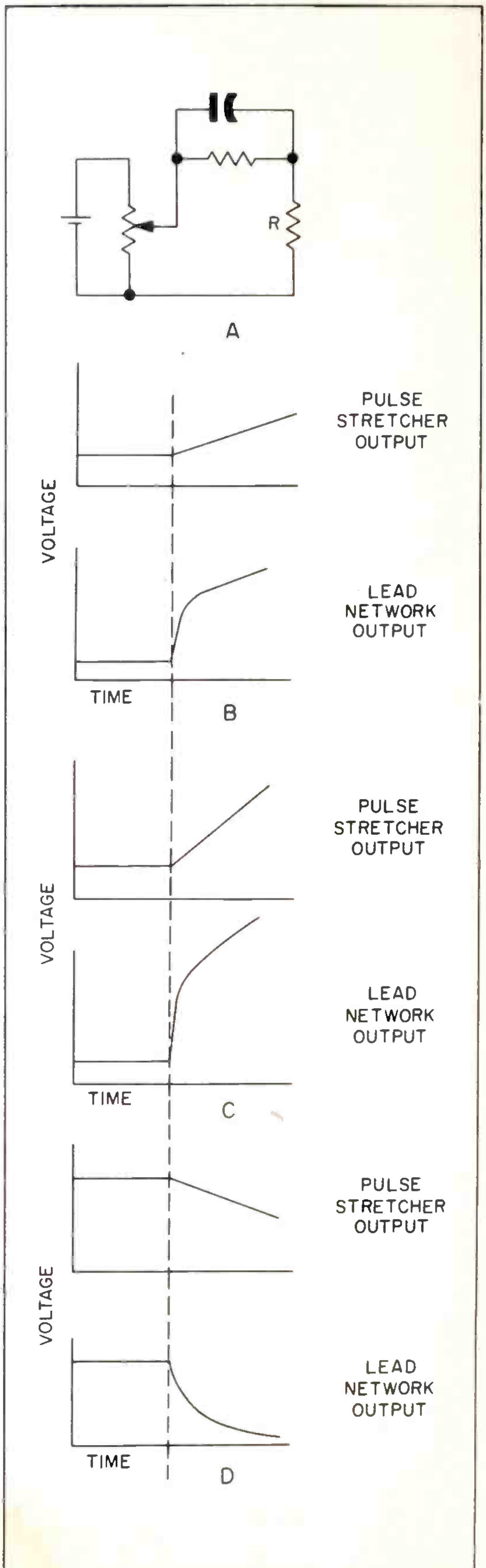

Figure 6. Operation of the Lead Circuit

motor from its initial position to the point where the scan-limit switch in the reference-voltage line is tripped. When this happens, the phase of the reference voltage applied to the tuning motor is reversed, and so is the direction of motor rotation. The motor is driven in the opposite direction until a cam drives the scan-limit switch back to its original position. This scanning action is illustrated in part A of figure 7. The frequency of the local oscillator is thus varied over a range of about 100 mc.

Locking In

Suppose that during the scanning time the transmitter is turned on, as illustrated in part B of figure 7, and that the local oscillator is to lock in at 30 mc. *above* the transmitter frequency. The condition necessary for lock-in for this instance is that the local-oscillator frequency, while scanning, approach the desired frequency from the low-frequency side. If the local oscillator is to lock in 30 mc. *below* the transmitter frequency, however, it must approach from above the desired operating frequency, as illustrated in part C of figure 7.

Locking in takes place as follows: The local oscillator approaches the correct frequency so as to first produce an i.f. *less* than 30 mc. The subsequent action is illustrated in figure 8, which shows the response curve of the discriminator. As the proper frequency is approached, relay K_1 de-energizes because the output of the upper pulse stretcher (refer to figure 4) is increasing and causes the current through d-c amplifier V_3 to increase. This increases the voltage across

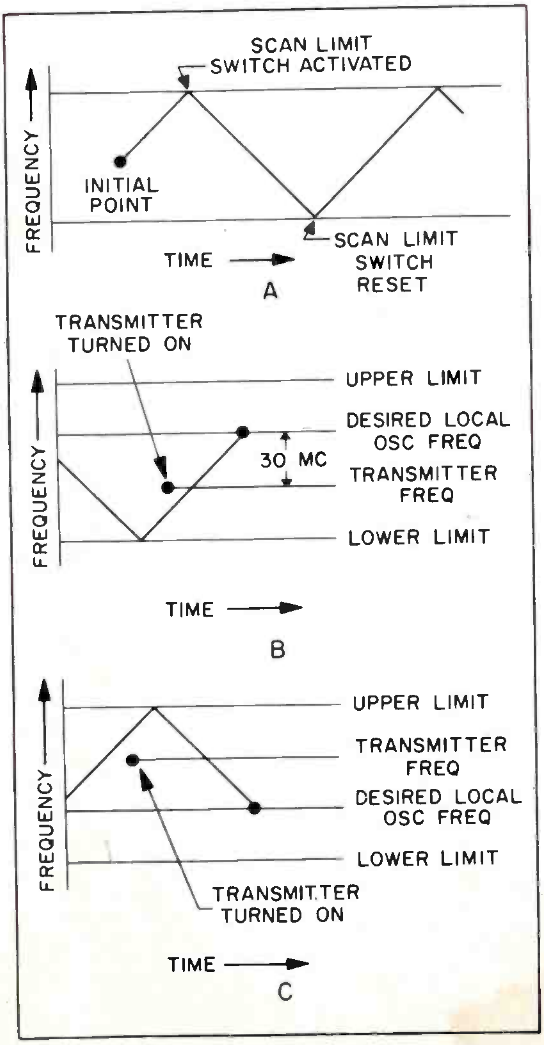

Figure 7. Action of Lighthouse-Tube Search Circuit

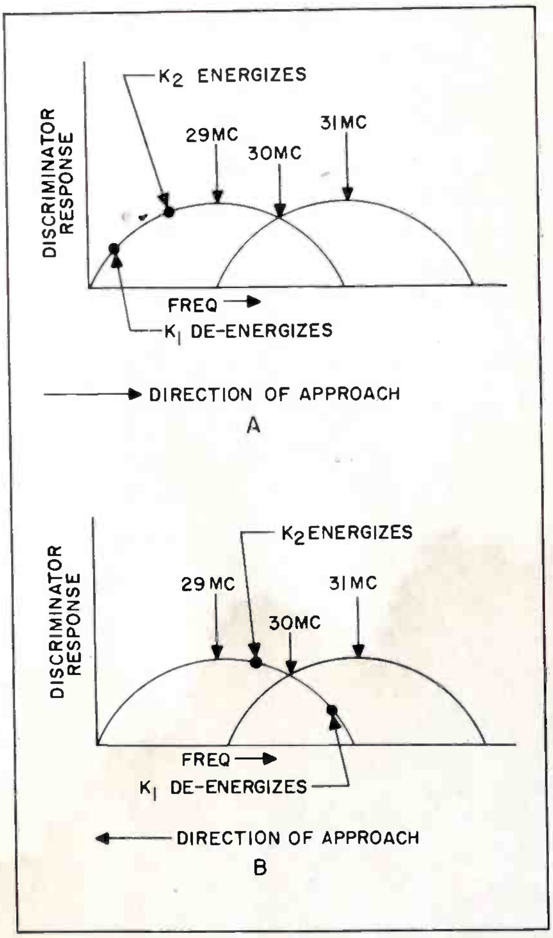

Figure 8. Relay Operation in the Lighthouse A-F-C Circuit

cathode resistor R_5 and reduces the current through d-c amplifier V_4 until relay K_1 de-energizes. At this time the short circuit is removed from relay K_2, but the current through it is still insufficient to allow it to energize. Some time later it does energize, however, as shown in figure 8. When K_2 energizes, the balanced modulator is returned to a balanced condition, and the discriminator output drives the local oscillator to the correct frequency.

The lower pulse stretcher is connected to relay amplifier V_5, in order to give a reasonable range of control to the circuit. If only the upper pulse stretcher were connected, relay K_2 would again be de-energized at about 30 mc. Note that when K_2 energizes, it connects the plate of V_5 in parallel with the plate of V_6 and both plate currents flow through K_2. Consequently, K_2 is energized by the upper pulse stretcher output alone, and remains energized because the current through it remains nearly constant from 29 to 31 mc. This constant current is a result of the fact that as the current in V_5 increases, the current in V_6 decreases when approaching 30 mc.

For an approach from the opposite direction, the discriminator output drives the local oscillator away from, instead of toward, the correct frequency because the phase of the reference voltage is reversed. This condition is shown in part B of figure 8.

"What's Your Answer?"

In an attempt to simplify the explanation of the Miller effect of an amplifier, a member of the Headquarters Technical Staff developed the circuit shown in figure 1. Although its use helps him in explaining the Miller effect, he finds that explaining the simplified circuit itself has now become a problem.

In figure 1, V_1 and V_2 represent two cascaded grounded-cathode amplifiers, and R_1 and R_2 are their respective a-c plate resistances. R_3 is the grid-to-cathode impedance of V_2, while R_4 is the plate-to-grid impedance of V_2.

Figure 2 depicts an equivalent circuit of figure 1, wherein V_1 and R_1 are identical with V_1 and R_1 of figure 1 and R_x represents the loading effect of all of the other elements of figure 1. With regard to figure 2, what is the value of R_x, and what is its physical significance?

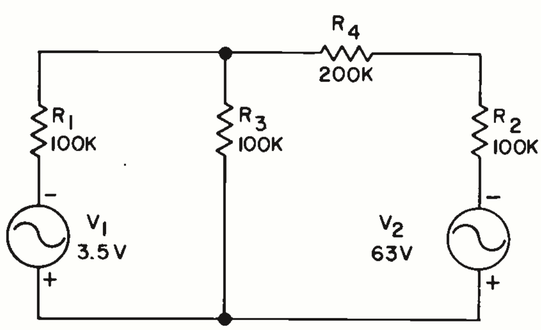

FIGURE 1

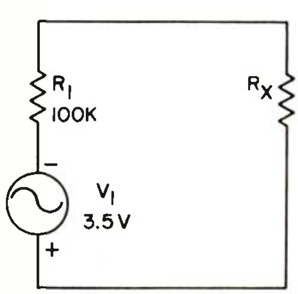

FIGURE 2

(Solution next issue . . .)

HYBRID JUNCTIONS
PART 2

by John E. Marchesano
Senior Engineer, Philco Research Division

This is the second and concluding article on the more common hybrid circuits used in radar applications.

THE HYBRID RING

The magic T has the disadvantage of small power-handling capability because of the matching devices that are required to prevent reflections at the junction. This characteristic is unimportant in low-power applications such as signal mixers, but prevents use of the magic T as a duplexer in high-power radar systems. For such applications the magic T is modified to form a ring. This hybrid ring, sometimes referred to as a "rat race," is a commonly used duplexer in modern radar systems. The functional location of the duplexer in a radar set is shown in figure 1. During the transmitting period, r-f energy is coupled from the transmitter to the antenna by means of the hybrid ring, and little energy reaches the receiver. During the reception period, the hybrid ring couples the received signals to the receiver and effectively prevents these signals from reaching the transmitter. A hybrid ring is shown in figure 2.

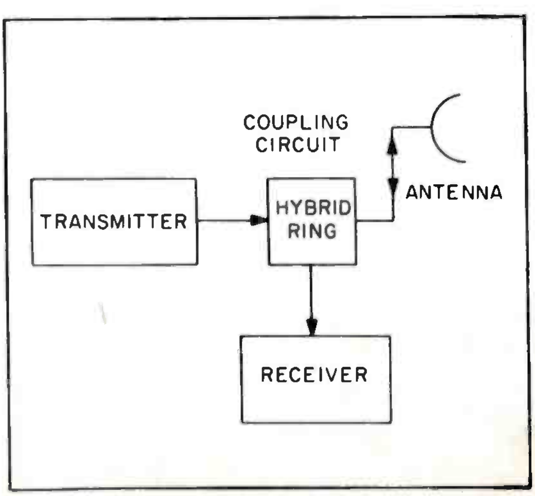

Figure 1. Position of Hybrid Ring Coupling Circuit in a Radar Set

The ring is constructed of rectangular waveguide. The waveguide arms designated as E, F, G, and H in figure 2 make E-type T junctions where they join the guide making up the ring. The two main functions of the ring are to couple the arms together in the desired manner, and to effect an impedance match for the arms. The first of these functions determines the circumference of the ring and the placement of the arms as indicated in figure 3, and the second determines the dimensions of the ring guide.

THE HYBRID RING DUPLEXER

Figure 3 shows a top view of the hybrid ring duplexer with the most important dimensions given in terms of wavelengths. The F and H arms contain TR tubes at the positions indicated. The gaps in these tubes ionize when sufficient r-f energy appears across them. When ionized, each TR tube effectively places a short circuit across its associated waveguide, and thus prevents r-f energy above a predetermined level from entering the receiver.

Operation During Transmission

The r-f energy from the transmitter must be coupled by the hybrid ring to the antenna, but not to the receiver. To understand this operation, first consider the situation which would result if arms H and F of figure 3 were not present. Figure 4 shows the circuit containing only arms G and E. The arrows across the waveguide in figure 4 represent the E field for one instant of time. Actually the wave travels around the ring with a velocity nearly that of light. Since the arms make E-type T junctions with the ring, energy entering the ring from arm

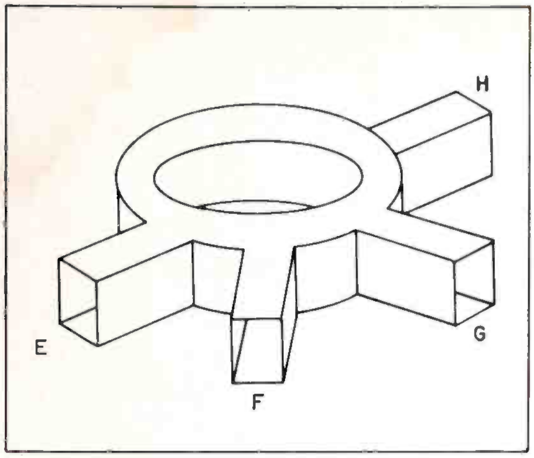

Figure 2. The Hybrid Ring

G will divide equally in both directions, but with opposite phases, as shown in figure 4. The energy traveling in a clockwise direction from arm G to arm E travels one-half wavelength before arriving at the E-arm junction. The energy traveling in a counterclockwise direction must cover a full wavelength before arriving at the same point. Since the total path difference is one-half wavelength, and the waves were out of phase at the G-arm junction, they will arrive at the E-arm junction in phase as indicated. This situation, however, will not produce a difference of potential across the

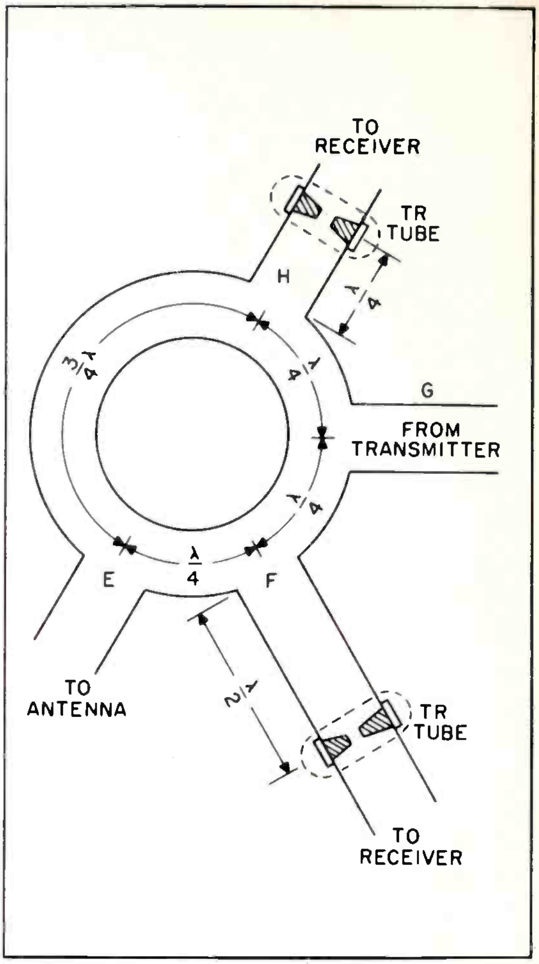

Figure 3. Top View Representation of Hybrid Ring Circuit

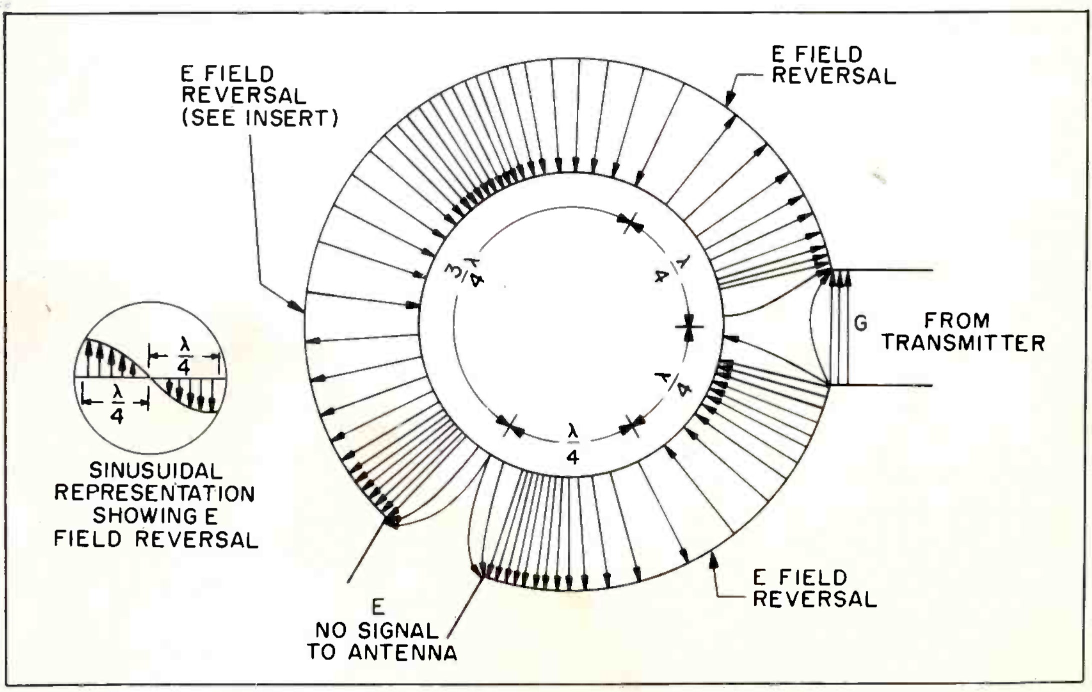

Figure 4. E Fields in Hybrid Ring with Only Arms G and E Shown

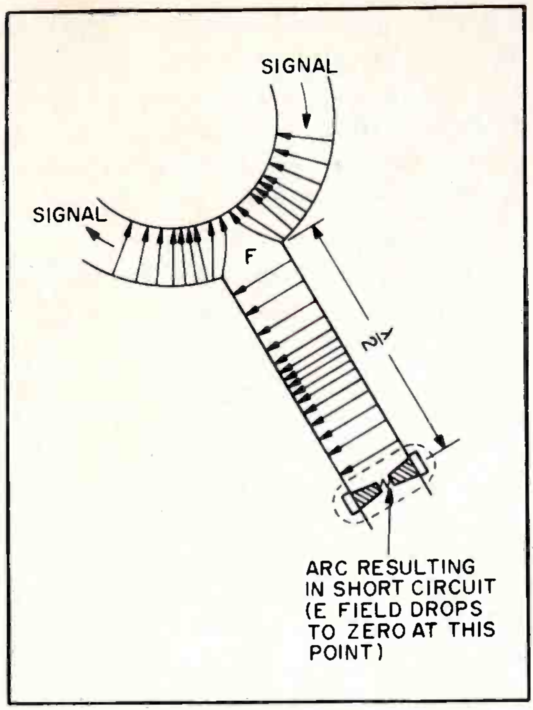

Figure 5. E Fields in Arm F When TR Tube Fires

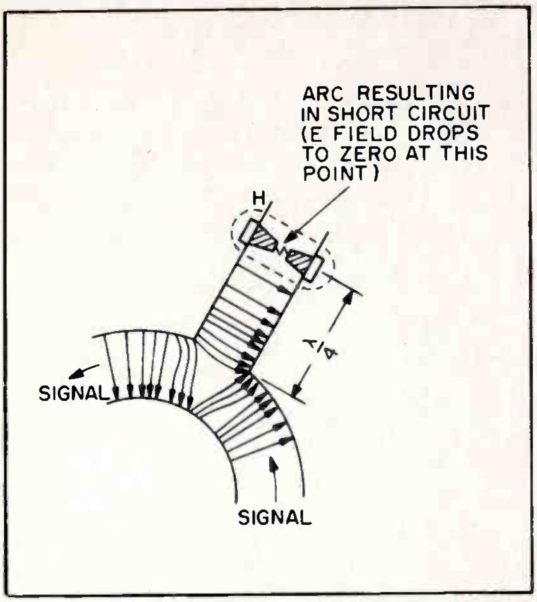

Figure 6. E Fields in Arm H When TR Tube Fires

E-arm guide; consequently, no r-f energy will be propagated along this guide.

It is apparent that a phase reversal of one of the signals traveling from arm G to arm E would cause the signals to arrive at the E-arm junction out of phase, which is the condition required to provide signal coupling. Consider the effect of adding arm F with a TR tube located one-half wavelength from its junction with the ring (see figure 5). When sufficient r-f energy arrives at the TR gap, the gap will ionize, placing a short circuit across the waveguide at this point. A short circuit will also appear across arm F at the T junction one-half wavelength away. As a result, the wave arriving at the F-arm junction will not see a discontinuity, and will proceed past the junction without a phase reversal. The necessary phase reversal is obtained through the addition of arm H. Figure 6 shows the electric field in the vicinity of the H arm. Here the TR tube is located one-fourth wavelength away from the T junction. When this TR tube is ionized by the r-f energy traveling in a counterclockwise direction around the ring, an open circuit is presented to the wave at this junction. As a result there will be a reversal of phase, as shown in figure 6. The necessary condition for coupling from arm G to arm E of the hybrid ring is now satisfied. The firing of the TR tubes in the F and H arms prevents additional r-f energy from reaching the crystal mixers of the receiver. Figure 7 shows the E field in the complete hybrid ring for an instant during the transmission period.

Operation During Reception

For reception of signals, it is necessary to couple arm E to arms F and H, but not to arm G, where energy would be lost in the transmitter. Shortly after the end of the transmitter pulse, the TR tubes de-ionize. The received signals do not have sufficient energy to ionize them, so the tubes remain de-ionized throughout the reception period. Figure 8 shows the configuration of the instantaneous E field for that portion of the received signal which is traveling in a clockwise direction around the ring. It changes polarity in the ring at the H-arm junction, but continues through the H arm to the receiver as indicated. The other half of the received signal traveling in a counterclockwise direction is

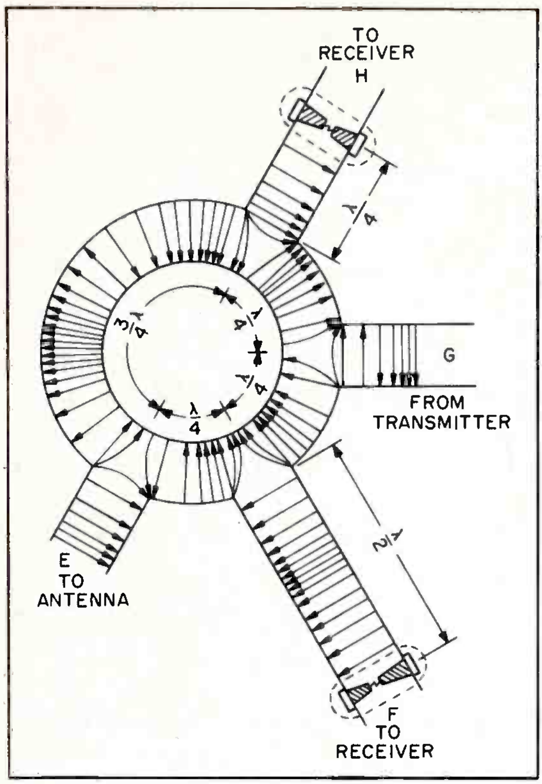

Figure 7. E Fields in Hybrid Ring with Signal from Transmitter

reversed in phase at the F-arm junction, as shown in figure 9. As a result, the situation is the same as when the H and F arms were omitted. There is no coupling between the E and G arms because the fields are in phase opposition at the G-arm junction.

A Three-Ring Duplexer Circuit

The primary disadvantage of older types of TR devices is the crystal burn-out caused by the leakage of r-f energy through the TR section. This leakage occurs because of the relatively high energy level required to ionize the TR gap. The next duplexer to be described more completely isolates the receiver from the transmitter, and allows for balanced mixing for the suppression of local-oscillator noise.

Three rings identical to that just discussed are connected in the manner shown in figure 10. The rings are numbered and the arms lettered to facilitate discussion. The long solid lines with

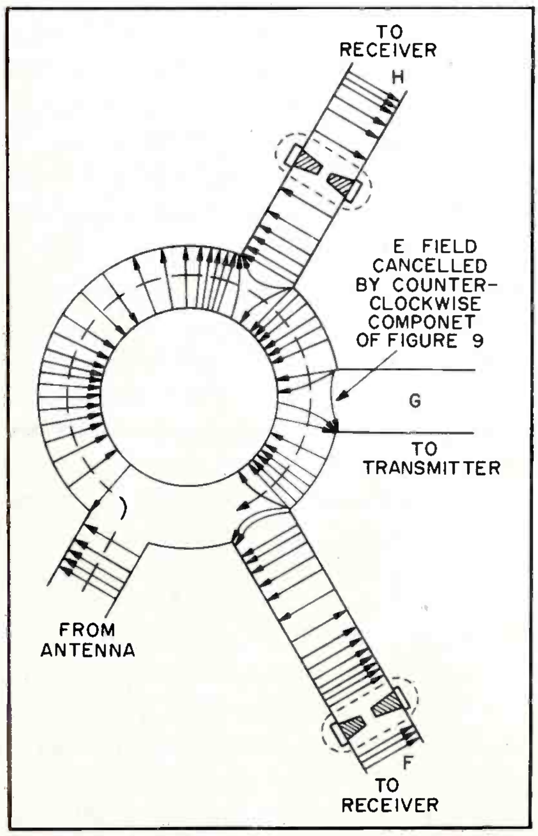

Figure 8. Clockwise Component of E Field Due to Echo Signal

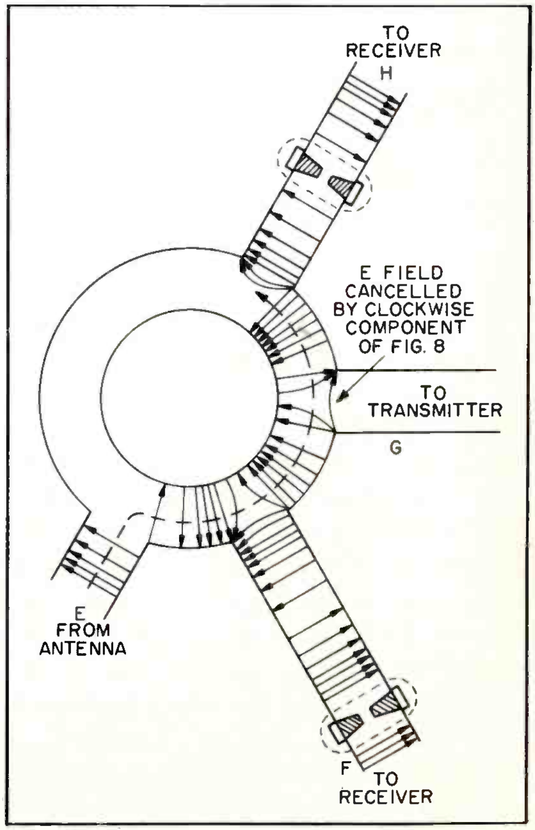

Figure 9. Counterclockwise Component of E Field Due to Echo Signal

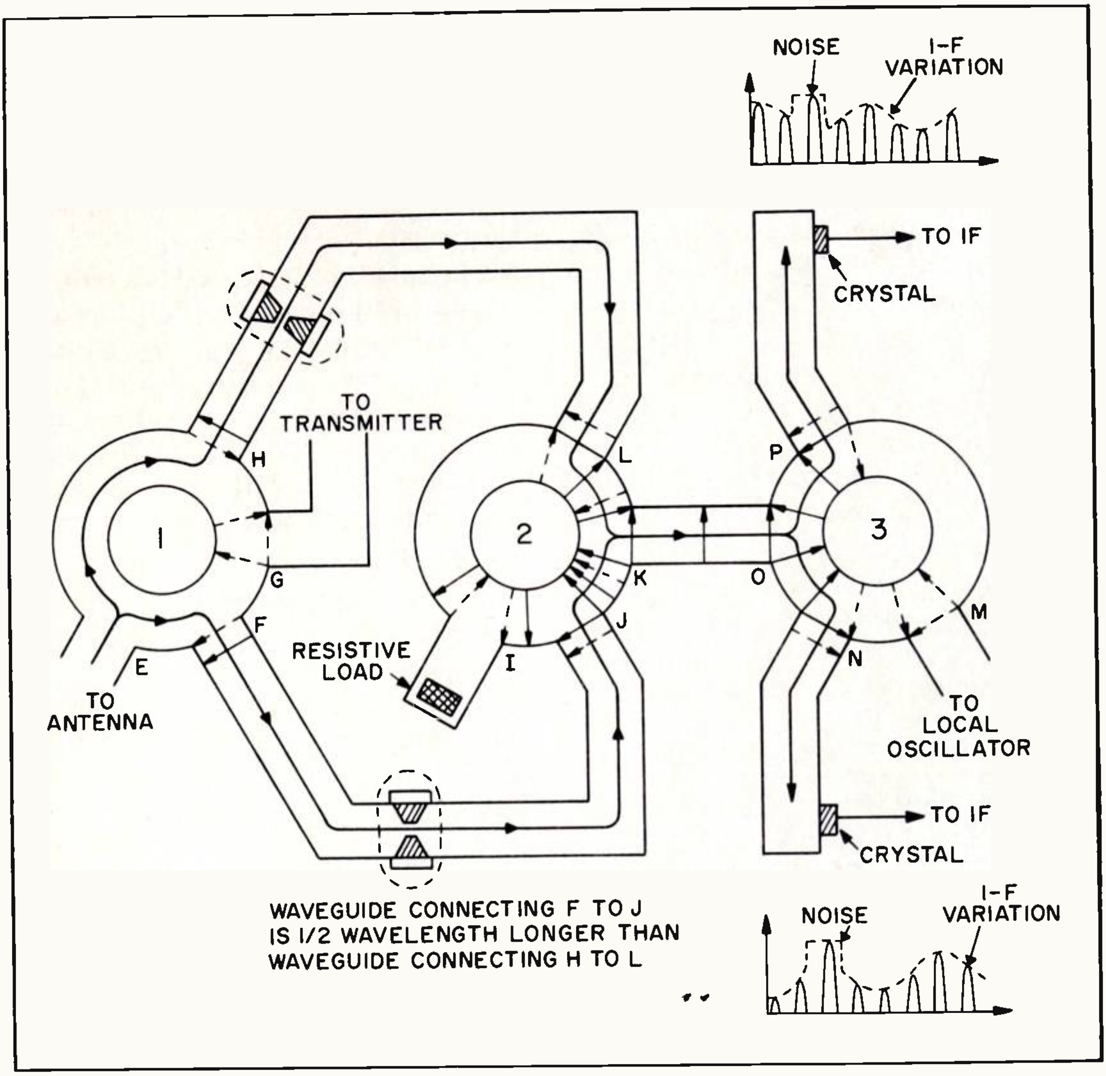

Figure 10. Three-Ring Hybrid Coupling Circuit

arrowheads indicate the paths taken by the received signal, and the solid E-field arrows represent the instantaneous phase of the received signal. The dotted E-field arrows in rings 1 and 2 represent the phase of the transmitter leakage signal, and in ring 3 represent the phase of the local-oscillator signal.

During the transmission period, the TR tubes ionize and isolate rings 2 and 3 from ring 1 and the transmitter. The operation of ring 1 is the same as that of the hybrid ring described previously.

During the reception period, energy is coupled from arm E to arms F and H of ring 1, but not to the transmitter arm G. Because of the phase reversal of one of the waves at the E-arm junction, and the one-half wavelength difference in their paths, the clockwise and counterclockwise waves leave the F-arm and H-arm junctions of ring 1 with the same phase. The addition of an extra one-half wavelength section of guide between arms F and J causes the signals arriving at the J-arm and L-arm junctions of ring 2 to again be in phase opposition. Since these junctions are equidistant from the K-arm junction, the E fields will be out of phase at the latter junction, precisely the condition required for propagation of the wave through the K-O arm. Arms L and J will not couple received energy to the I arm (which is used for the suppression of transmitted energy) of ring 2 because of the one-half wavelength difference in path length. Upon the arrival of the wave at the O-arm junction, another phase reversal occurs, so that the outputs of the N and P arms, which are both one-half wavelength from the O arm, are out of phase. Thus the phase relationship is correct for balanced-mixer operation. The local-oscillator signal enters ring 3 through arm M, and divides into two signals of opposite phase. Since the difference in path lengths to the N-arm and P-arm junctions is one-half wavelength, the local-oscillator signals will be of the required phase at these points. The outputs of the crystals are shown in figure 10. The balanced mixer operates in the manner described previously.

The path taken by the transmitter leakage pulse is shown in figure 10, where the dotted arrows represent the electric field of this signal. The transmitter signal divides into two signals of opposite phase at the G-arm junction of ring 1 as indicated. Since the path lengths to the J-arm and L-arm junctions of ring 2 differ by one-half wavelength, the transmitter signals are in phase at these points as well as at the K-arm junction of ring 2. As a consequence, none of the transmitter signal travels through arm K. The I arm, containing the resistive load, is one-fourth wavelength away from the J-arm junction, and three-fourths wavelength away from the L-arm junction. As a result, the transmitter signals are in phase opposition at the I-arm junction, the condition required for coupling to this arm. The resistive load then absorbs whatever r-f energy has leaked through the TR tube from the transmitter. For the reasons just discussed, none of the local-oscillator signal will be coupled to the O arm of ring 3. From the preceding discussion, it is apparent that the three-ring duplexer performs the usual TR functions very efficiently, and at the same time accommodates a balanced-mixing arrangement.

REFERENCES

Many other combinations of hybrid rings and junctions are possible and can be found in present day radar equipments. The following list of references may be consulted for additional information on the operation and construction of hybrid circuits.

1. Vol. 8, Radiation Laboratory Series, McGraw-Hill Book Company, Inc., 1948.
2. Vol. 10, Radiation Laboratory Series, McGraw-Hill Book Company, Inc., 1948.
3. Vol. 11, Radiation Laboratory Series, McGraw-Hill Book Company, Inc., 1948.
4. Vol. 14, Radiation Laboratory Series, McGraw-Hill Book Company, Inc., 1948.
5. Microwave and Radar Electronics, Ernest C. Pollard and Julian M. Sturtevant, John Wiley and Sons, Inc., 1948.

A TUNING PROCEDURE FOR AN/FPS-3 RADARS

The following tuning procedure has been used by the author and found to be especially useful at locations where ground clutter obscures the ring time of a radar set. The procedure has been in use for a period of about one year, and has proved to be beneficial to the maintenance of high radar system performance.

THE EQUIPMENT NEEDED to perform the tuning procedure is normally issued with the radar set, and consists of the following:

One R-F Signal Generator TS-419
One Crystal Detector UG-199/U
Three test cables

Procedure:

1. Connect the output of the crystal detector to the input of the radar test scope located in the radar maintenance room. The crystal detector should be located as near to the signal generator as possible.
2. Connect a coaxial T connector to the input of the crystal detector, and connect the output of the signal generator to one side of the T connector. Connect the directional coupler of the radar set to the remaining connection of the T connector.
3. Apply power to the TS-419, and allow it to warm up for about 20 minutes. Tune the signal generator approximately to the transmitter frequency, and make the proper adjustments of the zero set and power output controls of the signal generator.
4. Set the attenuator control of the signal generator to the zero db position. Place the test set in pulse operation, and set the pulse width control of the signal generator to equal the pulse width of the radar set.
5. Place the radar test scope video selector switch to the external video position, and adjust the scope to obtain a short sweep. A video presentation of the radar transmitter pulse should now be observed. By means of the delay control on the signal generator, superimpose the signal generator pulse on the radar transmitter pulse on the test scope.
6. With the two signals superimposed, tune the signal generator until a zero beat is observed on the test scope. The frequency of the signal generator output is now equal to that of the radar transmitter.
7. Disconnect the setup and connect the signal generator directly to the radar directional coupler. Check the signal generator pulse to make sure that it is not being limited in the receiver. Increase the attenuation of the signal-generator signal until limiting no longer occurs.
8. The radar set may now be tuned to maximize the signal generator pulse, which can easily be delayed to exceed the range of normal ground clutter.

Donald E. Faulkner
Philco TechRep Field Engineer

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