# RADAR BEACONS

*Edited by*

**ARTHUR ROBERTS**

ASSOCIATE PROFESSOR OF PHYSICS
STATE UNIVERSITY OF IOWA

OFFICE OF SCIENTIFIC RESEARCH AND DEVELOPMENT
NATIONAL DEFENSE RESEARCH COMMITTEE

FIRST EDITION

RADIATION · LABORATORY · M · I · T · CAMBRIDGE · MASS · 1940 · 1945

*NEW YORK AND LONDON*
*McGRAW-HILL BOOK COMPANY, INC.*
1947

*MASSACHUSETTS INSTITUTE OF TECHNOLOGY*

RADIATION LABORATORY SERIES

LOUIS N. RIDENOUR, *Editor-in-Chief*

---

# RADAR BEACONS

## *MASSACHUSETTS INSTITUTE OF TECHNOLOGY*
# RADIATION LABORATORY SERIES

---

1. Radar System Engineering—*Ridenour*
2. Radar Aids to Navigation—*Hall*
3. Radar Beacons—*Roberts*
4. Loran—*Pierce, McKenzie, and Woodward*
5. Pulse Generators—*Glasoe and Lebacqz*
6. Microwave Magnetrons—*Collins*
7. Klystrons and Microwave Triodes—*Hamilton, Knipp, and Kuper*
8. Principles of Microwave Circuits—*Montgomery, Dicke, and Purcell*
9. Microwave Transmission Circuits—*Ragan*
10. Waveguide Handbook—*Marcuvitz*
11. Technique of Microwave Measurements—*Montgomery*
12. Microwave Antenna Theory and Design—*Silver*
13. Propagation of Short Radio Waves—*Kerr*
14. Microwave Duplexers—*Smullin and Montgomery*
15. Crystal Rectifiers—*Torrey and Whitmer*
16. Microwave Mixers—*Pound*
17. Components Handbook—*Blackburn*
18. Vacuum Tube Amplifiers—*Valley and Wallman*
19. Waveforms—*Chance, Hughes, MacNichol, Sayre, and Williams*
20. Electronic Time Measurements—*Chance, Hulsizer, MacNichol, and Williams*
21. Electronic Instruments—*Greenwood, Holdam, and MacRae*
22. Cathode Ray Tube Displays—*Soller, Starr, and Valley*
23. Microwave Receivers—*Van Voorhis*
24. Threshold Signals—*Lawson and Uhlenbeck*
25. Theory of Servomechanisms—*James, Nichols, and Phillips*
26. Radar Scanners and Radomes—*Cady, Karelitz, and Turner*
27. Computing Mechanisms and Linkages—*Svoboda*
28. Index—*Henney*

# RADAR BEACONS

*Edited by*

ARTHUR ROBERTS

ASSOCIATE PROFESSOR OF PHYSICS

STATE UNIVERSITY OF IOWA

OFFICE OF SCIENTIFIC RESEARCH AND DEVELOPMENT

NATIONAL DEFENSE RESEARCH COMMITTEE

FIRST EDITION

RADIATION · LABORATORY · M·I·T · CAMBRIDGE · MASS · 1940-1945

*NEW YORK AND LONDON*
*McGRAW-HILL BOOK COMPANY, INC.*
1947

RADAR BEACONS

PRINTED IN THE UNITED STATES OF AMERICA

THE MAPLE PRESS COMPANY, YORK, PA.

## *RADAR BEACONS*

### *EDITORIAL STAFF*

Louis A. Turner

Gordon C. Danielson
Peter A. DePaolo
Erwin R. Gaerttner
Ray G. Herb
Conrad L. Longmire
J. J. G. McCue
Marcus D. O'Day
Beuhring W. Pike
William M. Preston
Arthur Roberts

### *CONTRIBUTING AUTHORS*

Howland H. Bailey
Martin J. Cohen
Gordon C. Danielson
Peter A. DePaolo
Raymond Dickinson
William A. Downes
Erwin R. Gaerttner
Serge E. Golian
Gilbert O. Hall
Jesse R. Lien
Conrad L. Longmire
J. J. G. McCue
Raymond A. Minzner
Kenneth R. More
Marcus D. O'Day
Beuhring W. Pike
William M. Preston
John C. Reed, Jr.
Arthur Roberts
Theodore S. Saad
John H. Tinlot
Louis A. Turner
George Peter Wachtell
Frank P. Zaffarano

# Foreword

THE tremendous research and development effort that went into the development of radar and related techniques during World War II resulted not only in hundreds of radar sets for military (and some for possible peacetime) use but also in a great body of information and new techniques in the electronics and high-frequency fields. Because this basic material may be of great value to science and engineering, it seemed most important to publish it as soon as security permitted.

The Radiation Laboratory of MIT, which operated under the supervision of the National Defense Research Committee, undertook the great task of preparing these volumes. The work described herein, however, is the collective result of work done at many laboratories, Army, Navy, university, and industrial, both in this country and in England, Canada, and other Dominions.

The Radiation Laboratory, once its proposals were approved and finances provided by the Office of Scientific Research and Development, chose Louis N. Ridenour as Editor-in-Chief to lead and direct the entire project. An editorial staff was then selected of those best qualified for this type of task. Finally the authors for the various volumes or chapters or sections were chosen from among those experts who were intimately familiar with the various fields, and who were able and willing to write the summaries of them. This entire staff agreed to remain at work at MIT for six months or more after the work of the Radiation Laboratory was complete. These volumes stand as a monument to this group.

These volumes serve as a memorial to the unnamed hundreds and thousands of other scientists, engineers, and others who actually carried on the research, development, and engineering work the results of which are herein described. There were so many involved in this work and they worked so closely together even though often in widely separated laboratories that it is impossible to name or even to know those who contributed to a particular idea or development. Only certain ones who wrote reports or articles have even been mentioned. But to all those who contributed in any way to this great cooperative development enterprise, both in this country and in England, these volumes are dedicated.

L. A. DUBRIDGE.

# Preface

THIS book is about radar beacons. As far as the authors are aware, no other books on radar beacons have yet appeared. Because beacons constitute an important aspect of radar, an attempt will be made in this book to give a comprehensive survey of the present state of the beacon art.

Beacons were, at first, a minor and neglected aspect of the radar art. Their usefulness was recognized in nearly all laboratories working on radar, but only after the design of the radar was well advanced. Organized work on beacons at the Radiation Laboratory started only when the laboratory was in its second year; a separate beacon division was not organized until the summer of 1943. The influence of beacons was slow in making itself felt; nevertheless, by the end of the war, it had established itself firmly in the thinking of the designers of airborne and ground radar equipment and was beginning to be accepted for ship radar.

This survey is divided into four parts. Part I, Chaps. 1 to 6, discusses the nature of beacons and the principles according to which systems using beacons are designed. Part II, which includes Chaps. 7 to 16, is concerned with the design of beacons. Part III, Chaps. 17 to 19, takes up the design of beacon interrogators and gives examples of complete systems using beacons. Part IV, Chap. 20, covers the operation of beacon systems in the field.

This volume represents a summary of the efforts of Division 7 of the Radiation Laboratory, which was charged with work on radar beacons. In an attempt to give a complete picture of the beacon art, however, the authors have included much that did not originate at the Radiation Laboratory.

The plan of the book was drawn up by an editorial board before the personnel of the division was dispersed. Many members of the division were thus able to contribute to the book; and it could certainly never have been written without the contributions of so large and representative a group of the members of the division. Any omissions are due either to inadvertence or to residual military security.

Full acknowledgment to everyone who contributed to this book is impossible; apology must be made in advance for any omissions. First acknowledgment must go to L. A. Turner, who, as technical editor, guided the manuscript from the beginning. A special debt of gratitude is due to B. V. Bowden of the British Air Commission, whose thorough acquaintance with the subject and friendly interest in the manuscript were of inestimable value. The authors wish to express their appreciation to J. R. Feldmeier, S. A. Goudsmit, D. E. Kerr, and F. F. Rieke for their constructive comments and contributions to the manuscript, to the RCA-Victor Company for permission to describe Shoran, and to Stuart W. Seeley of that company for comments on its treatment.

Thanks are due to Constance R. Henderson, who, as production assistant, shepherded the manuscript and figures through the mazes of production; and to Nina M. Kropoff and Margaret Jordan for their secretarial assistance. Especial thanks are owed to Beka Doherty, to whose editorial efforts is due what degree of literacy the book may possess, and who aided immeasurably in its final organization and intelligibility.

ARTHUR ROBERTS.

CAMBRIDGE, MASS.,
*May*, 1946.

# *Contents*

## PART I. BASIC CONSIDERATIONS

## PART II. BEACON DESIGN

## PART III. INTERROGATOR AND SYSTEM DESIGN

## PART IV. BEACONS IN THE FIELD

# PART I

# BASIC CONSIDERATIONS

"They also serve who only stand and wait."

## CHAPTER 1

# THE USES OF BEACONS

By L. A. Turner and A. Roberts

## THE NATURE OF RADAR BEACONS

**1·1. The Echo and the Beacon.**—As is well known, a radar set operates by sending out powerful pulses of radio waves and then receiving the portions of the energy that are reflected back to it. The elapsed time between the emission of a pulse and the return of an echo is a measure of the distance to the reflecting target. Further, by means of suitable antenna systems, the radio energy is concentrated into narrow beams—in somewhat the same way that light is concentrated into a beam by a searchlight—so that the echoes are received only when the radar set is looking at the target. By proper coordination of the motion of the antenna and the sweep of an intensity-modulated beam of electrons of a cathode-ray tube, a plan view of the reflecting targets in the region of the radar set is traced out for the use of the operator.

The simple radar set is entirely adequate for many purposes. Often, however, echoes are too weak to be observed, as are those from a small airplane at a great distance, for example. Under other circumstances, strong echoes from numerous buildings, hills, or mountains mask weaker echoes from targets of greater interest. Furthermore, for the benefit of aircraft carrying radar sets, it is sometimes desired to mark some particular place on the ground that gives no distinguishing echo. In all such cases radar beacons find their use.

A radar beacon is a device that, upon reception of the original radar pulses, triggers its own transmitter to give a strong reply of one or more pulses independent of the possible radar echoes from the vicinity. The beacon, therefore, may be said to act as an amplifier of the echo. The beacon transmitter need not be very powerful to be able to give a reply much stronger than the echo from usual targets. Figure 1·1 shows the appearance of beacon replies on a plan-position indicator (PPI).

**1·2. What the Beacon Is and Does.**—A radar beacon is a device that is normally quiescent; it is passive. Without external stimulation, a radar beacon does nothing of any interest whatever. When a suitable pulsed signal reaches it, however, the beacon is actuated—the standard terminology is "triggered"—by the received signal, and emits a pulse or series of pulses. The process by which a radar set transmits a signal suitable for triggering the beacon is known as "interrogation"; the communication link that is thus established is called the "interrogation link." Any device used to interrogate a beacon may be designated as an "interrogator."

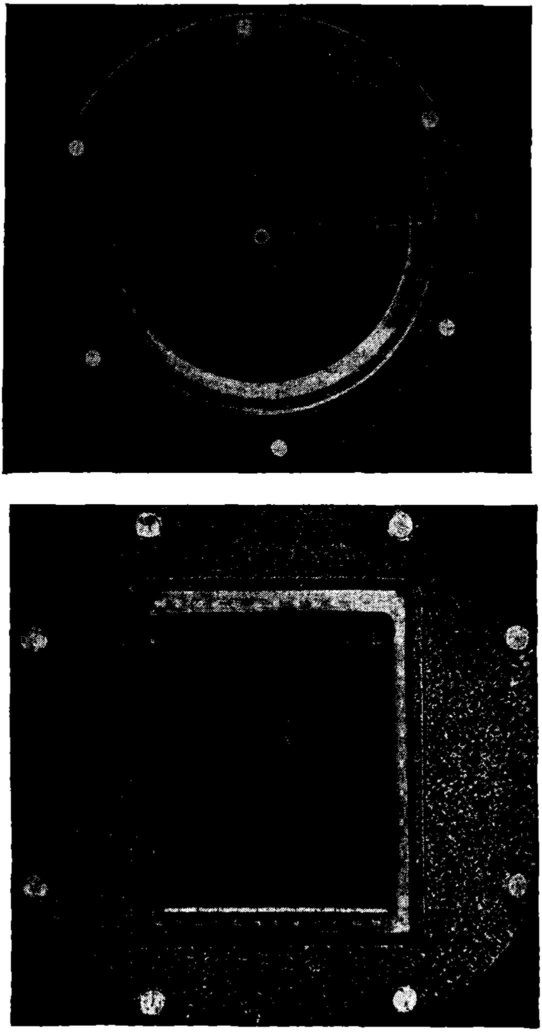


FIG. 1·1.—Beacon presentation on PPI and type *B* displays. Azimuth is read to the center of the reply, range to the nearest reply arc.

When the beacon is interrogated, it emits signals that are received by the interrogator. The independent communication link thus established is called the "response link"; the beacon is said to have "replied" or "responded" to the interrogation. In general, then, the operation of a beacon system may be visualized as shown in Fig. 1·2.

A ground beacon that is interrogated by suitable pulses from an airborne interrogator has been chosen arbitrarily for an illustration. The interrogator in the aircraft emits suitable interrogation pulses that reach the beacon receiving antenna. These signals actuate the beacon transmitter, which sends out a pulsed reply; the reply is then detected by a receiver associated with the interrogator. Although the response is like an echo, it differs from one in several significant respects. The strength of the response is independent of the intensity of the interrogating signal, provided only that the interrogating signal exceeds a threshold value of intensity. The response frequency is independent of the interrogation frequency, and in practice is usually different from it. The response signal pulse may differ from the interrogation signal in form. It may even consist of more than one pulse, the duration and spacing of the pulses being arbitrary. Finally, an unavoidable delay of a few microseconds' duration is introduced between the reception of the interrogating signal and the emission of the reply. This delay is usually a fixed amount for any particular beacon; the range of the beacon may be

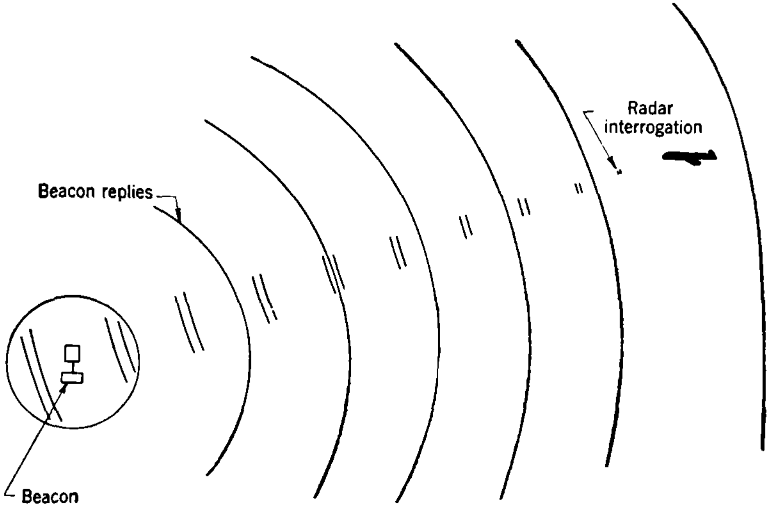


FIG. 1·2.—The operation of an airborne-interrogator–ground-beacon system. When the directional beam of the interrogator sweeps over the beacon, omnidirectional replies are emitted.

determined accurately by taking the known value of the delay into account.

**1·3. Terminology.**—Radar beacons have been called, variously, "beacons," "transponders," "responder beacons," and "racons." There are no generally recognized differences of meaning among these different terms.

A device used to interrogate a beacon is generally called an "interrogator." The interrogator may be a radar set. It may be a device designed especially to interrogate beacons; in this case it is called an "interrogator-responsor." The British have designed auxiliary equipment to be used with radar sets that do not themselves interrogate beacons. Such an auxiliary equipment they call an "inquisitor."[1]

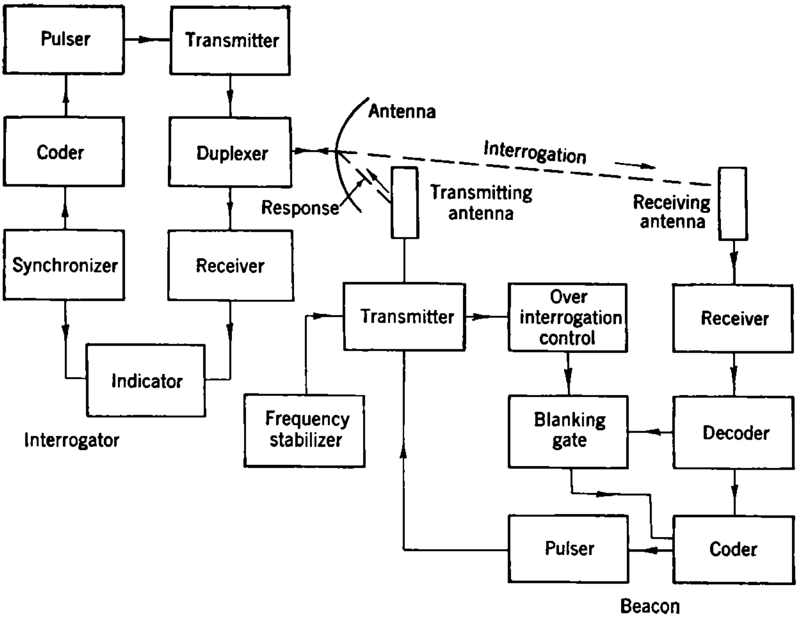


FIG. 1·3.—Block diagram of a system using a radar beacon.

The provision of a special character to the pulsed signals of either the interrogation or the reply link is called "coding." A device that sets up such a coded signal is called a "coder"; a device that deciphers the code at the other end of the link is called a "decoder" (or, sometimes, a "discriminator").

---

[1] The most-used example of this type was named "Lucero," after a particularly interesting sadist of the Spanish Inquisition.

**1·4. Block Diagram of a Beacon System.**—Figure 1·3 shows a block diagram of a system consisting of a beacon and a radar set used as an interrogator. In this diagram many of the important features of beacon-system design are illustrated.

The complete cycle of operation starts with the generation of a trigger pulse by the synchronizer in the radar set. This pulse initiates the transmission of an interrogating signal by the radar set, and also starts the cathode-ray-tube sweep on the indicator. The transmitted signal is received by the beacon receiving antenna, which passes it to the beacon receiver. The beacon receiver puts out a video signal to the decoder, if there is one. The decoder examines the video signal to see if it conforms to an acceptable code; if it does not, it is rejected. If it is accepted, a trigger pulse is passed to the blanking gate and to the coder. The blanking gate generates a pulse that prevents the coder from responding to any further triggers for a time sufficient to permit the complete response code to be generated. This time is generally from about 50 to 150 $\mu$sec. The coded response signal is supplied to the modulator, which raises it to a power level suitable for pulsing the transmitter and, in addition, shapes the pulses properly. The transmitter, often controlled in frequency by a stabilizing device, supplies r-f pulses to the transmitting antenna.

The response of the beacon is received by the radar antenna, detected by the receiver, and displayed on the indicator.

Some of the features in the system illustrated may be left out. The overinterrogation control has the purpose of preventing transmitter overload by lengthening the pulse supplied by the blanking gate when the number of interrogations becomes too large. A frequency stabilizer is not always required; neither are the coder and decoder. The beacon may employ a duplexer and use only one antenna instead of separate antennas.

The interrogator may have separate transmitting and receiving antennas. It may or may not have a coder to impart a special character to the interrogating pulses, and it may or may not include a decoder so that only suitably coded reply signals are displayed on the indicator. The desirability and use of these and other optional features will be discussed in succeeding chapters.

## HOW BEACONS HAVE BEEN USED

The radar beacon was invented, to the best of our knowledge, in 1939 by a group[1] at the Bawdsey Research Station of the Air Ministry. It came into being in response to a military need, and its invention was not

---

[1] The group included F. C. Williams, R. H. A. Carter, D. H. Preist, and R. W. Taylor.

made public for reasons of security. During the war radar beacons were used by the Germans and by the Japanese as well as by the Allies. It is not known with certainty at present whether the enemy engineers invented beacons independently or whether they learned of them from the original British use. The former alternative seems more likely, because the idea is not especially difficult to come by. Certainly beacons were invented independently in the United States. In any case, the German beacons were found to differ significantly from the first British beacons and to be superior in some respects.

Because the invention, in all these countries, was veiled by military security, there were no prewar nonmilitary uses of beacons. A summary of their past use, therefore, is restricted to military uses. This is merely a historical accident. There are many peacetime applications of beacons; their past use for military purposes exclusively should not prevent the exploration of more pacific applications.

**1·5. IFF and GCI.**—The initial purpose of beacons, the one that led to their greatest numerical use by far, was for IFF (Identification of Friend or Foe).

In the early part of the war, before the Allies had organized against the German power that had pushed them off the Continent, their strategy had to be primarily defensive. The Royal Air Force was numerically weak; fortunately it was superior to the Luftwaffe in quality. The function of the radar network ringing the British coast was to give warning of the approach of hostile aircraft. A radar echo from an aircraft carries no national insignia, however; it cannot reveal hostility. The decision to install radar beacons on all British aircraft—and later on all Allied combat aircraft—was intended to permit the differentiation of echoes into two sorts: friendly and hostile. Knowledge of the nature of approaching aircraft made it unnecessary to investigate all approaching aircraft indiscriminately, and thus conserved the strength of the interceptor groups. In addition, night fighters equipped with aircraft-interception (AI) radar were greatly aided in their task by being able to disregard friendly airplanes and to concentrate on attacking hostile ones.

It was soon discovered that beacons on night fighters could be used for offensive purposes. Since the ground radar stations were able to spot both enemy and friendly aircraft, they could give directions to friendly night fighters to help them in their attacks against hostile bombers. A technique for ground-controlled interception, GCI, was developed, in which the IFF beacon played an essential role. The beacon in GCI was not used solely for identification of an airplane as friend or foe, but also for identification as a particular friend—in other words, for identification of a particular night fighter. The IFF beacon was used for this purpose throughout the defensive phase of the war. Later, with the advent of

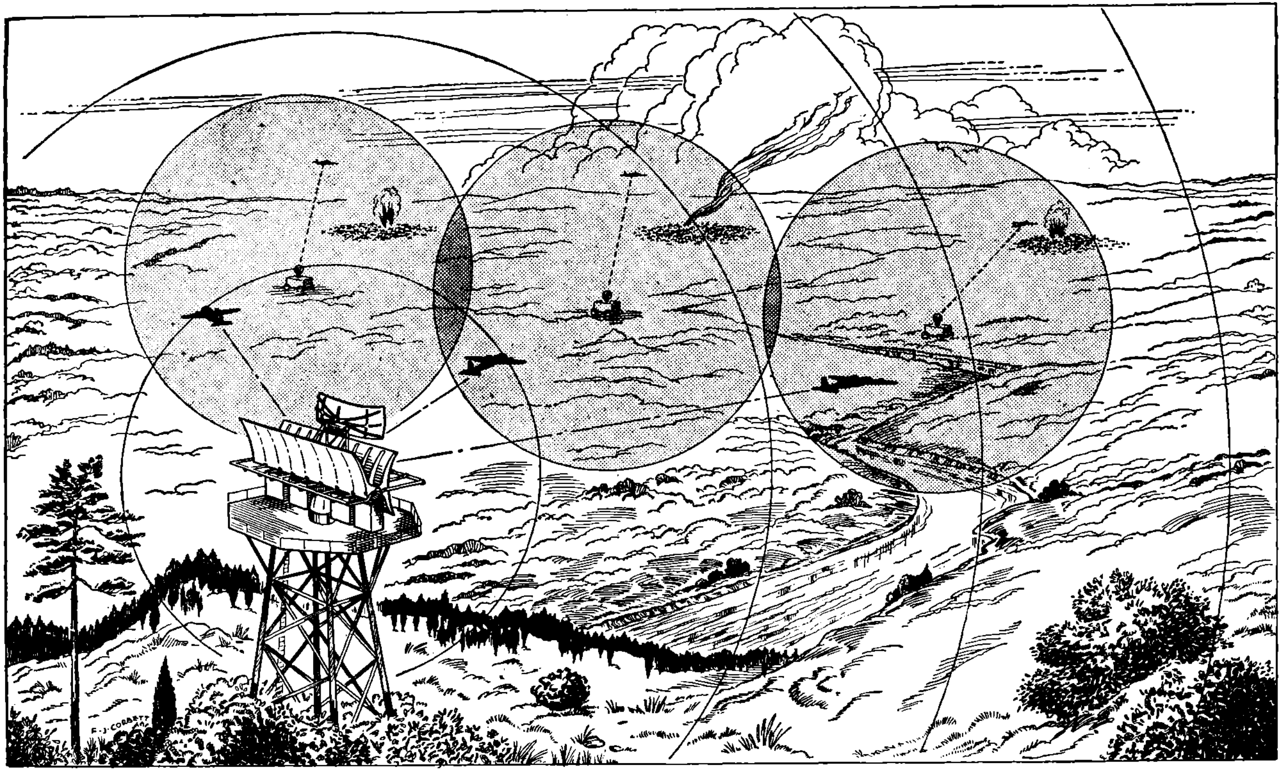


FIG. 1·4.—The use of beacons for ground control of aircraft. Aircraft are guided by a ground radar set to a point within range of a tactical control set like the SCR-584.

microwave ground-control stations, special beacons were developed for all ground-control purposes including purposes other than interception. Figure 1·4 shows how such control was carried out.

**1·6. Ground Beacons for Air Navigation.**—The antisubmarine campaign, important throughout the war, was especially significant in its early months. British Coastal Command ASV (air-to-surface-vessel) radar-equipped aircraft played an important part in the campaign. To assist these airplanes in homing to their bases and in making rendezvous with Allied convoys, beacons were installed at bases and on ships. These beacons served for both identification and navigation. Similarly, beacons were used on the ground to assist radar-equipped night fighters both in patrol and in homing to their bases.

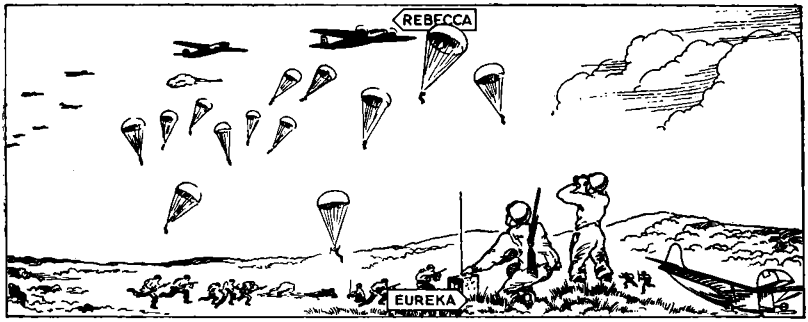


Fig. 1·5.—The use of Rebecca-Eureka in paratroop operations. Aircraft equipped with Rebecca home on Eureka beacons on the ground; minimum dispersion of the forces on the ground is thus assured.

The recognized value of these beacons working with the low-frequency AI and ASV radar led to the development of analogous beacons for use with microwave AI and ASV. This was the initial step in the development of microwave beacons in the United States, which was under way in the Radiation Laboratory by the end of 1941.

By the end of the war a network of navigational beacons at both high and low frequencies extended over all continents and constituted an important aid to navigation of radar-equipped aircraft.

**1·7. Lightweight Ground Beacons for Tactical Uses.** *Rebecca-Eureka.*—The first, and easily the most important, tactical use of lightweight ground beacons was their employment in the Rebecca-Eureka system. This system included a series of airborne interrogator-responsors called "Rebecca" and a series of ground beacons called "Eureka." Originally designed at Telecommunications Research Establishment, (TRE), for British use, the Rebecca-Eureka system was later adopted by the United States Army and used in many operations.

In the most frequent use of the Rebecca-Eureka system, Eureka beacons were carried down to the ground by pathfinder groups of paratroopers and set up immediately. Airplanes carrying the main paratroop force were equipped with Rebecca. By homing on the Eureka beacons, they were able to carry the paratroops to the correct dropping zones and to discharge them in well-concentrated groups that could be organized on the ground in a minimum of time (see Fig. 1·5). Rebecca-Eureka was used in this way in Sicily, Italy, Normandy, southern France, in the Burma-China Theater, and elsewhere. After initial operational difficulties were surmounted, the system became the main reliance of Troop Carrier Command for such operations.

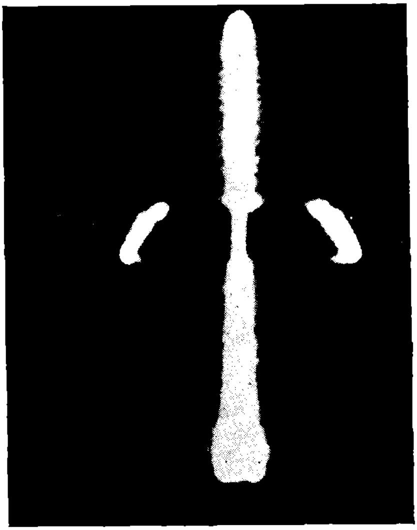


FIG. 1·6.—Type L presentation. Azimuth determinations are made by rotating two antenna arrays until the deflections to left and right are equal; the antenna crossover vector is then pointing at the target.

Other uses of Rebecca-Eureka suggested themselves. The original purpose of the system was to facilitate rendezvous with secret agents operating behind enemy lines. Eurekas were supplied to partisan and guerilla forces in northern Italy, France, and Yugoslavia. Operated behind enemy lines, they made possible the delivery of supplies and personnel by aircraft, requiring previous agreement only on the general area within which a rendezvous was to be effected. The hand-operated reply coding of the Eureka identified the particular beacon and provided a valuable safeguard against enemy deception by use of captured beacons.

The Rebecca-Eureka system operated near 200 Mc/sec and used lobe-switched antennas for azimuth determination and type L presentation (see Fig. 1·6).

*Lightweight Microwave Ground Beacons.*—Lightweight ground beacons were developed at 10 cm and 3 cm for use with microwave airborne radar. Of these the 10-cm beacons were used as paratroop beacons with the SCR-717 as an interrogator (as Rebecca-Eureka was used), and also to mark traffic lanes for units of the Troop Carrier Command in the Normandy landings. They were used also as portable navigational beacons at advanced air bases.

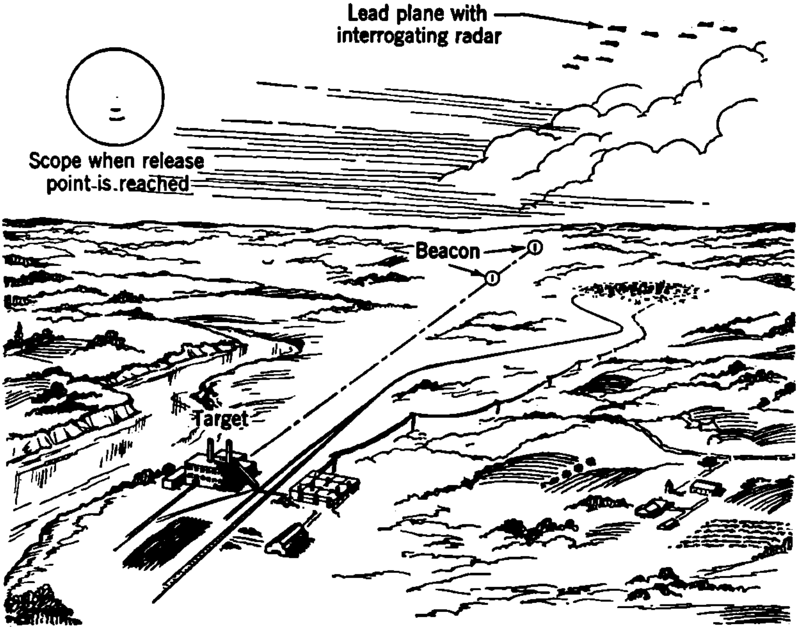


FIG. 1·7.—Target designation by portable beacons. A formation of aircraft led by an aircraft equipped with an interrogating radar can drop bombs on a point designated by one or more beacons some miles away. The beacon replies shown here are uncoded. The proper release point is reached when the aircraft are in line with the two beacons and some distance beyond them.

The lightweight 3-cm beacons that were in production at the end of the war were intended for various purposes. Among these were the designation of targets to tactical bombers and to ships off shore, and the establishment of precise coordinates (by the *H*-system, see Sec. 1·10) whereby the initial contingent of pathfinder paratroops might be dropped accurately at any desired location. (See Figs. 1·7 and 1·8.)

**1·8. Airborne Beacons.**—Beacons installed in aircraft were used during the war for IFF and GCI, as we have already noted. Among the other uses found for airborne beacons were ground control for purposes other than interception, and aircraft rendezvous.

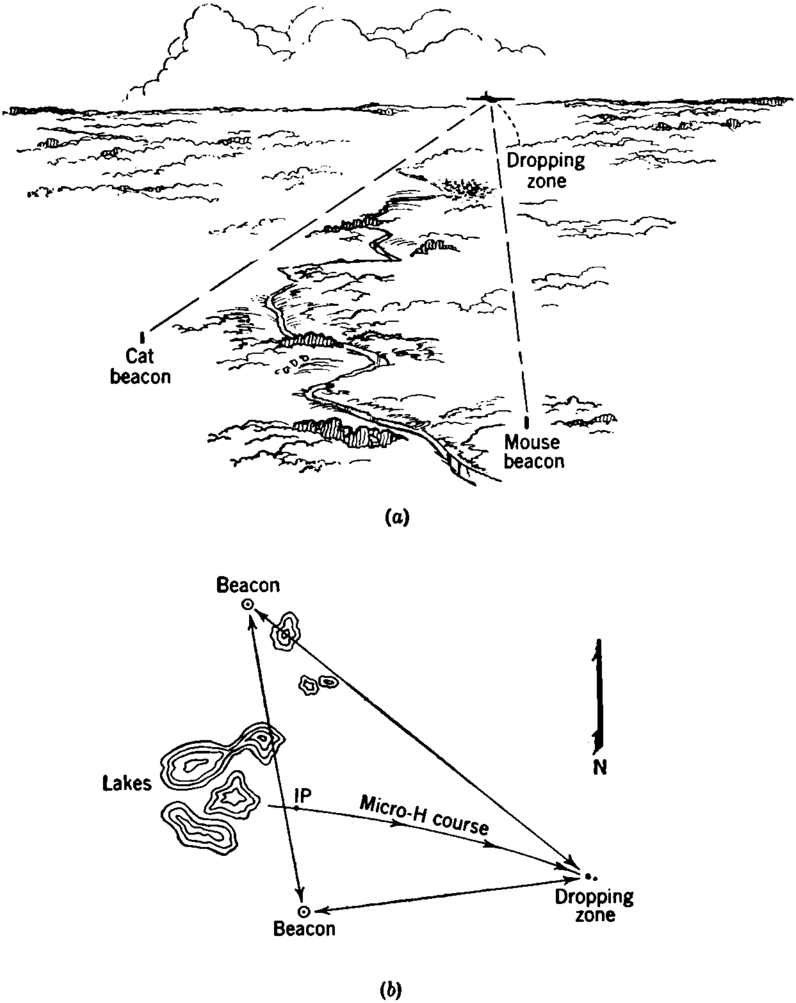


FIG. 1·8.—The use of portable beacons to designate dropping zones for paratroop pathfinders. A "cat-mouse" course (a) or a Micro-H hyperbolic course (b) can be flown (see Sec. 1·10). The beacons are located in friendly territory.

*Air-to-air Rendezvous.*—It is clear that an aircraft equipped with an interrogator may home on a beacon either at an airport or in another aircraft.

In sea-search operations, when a search plane had located an object of interest, such as a submarine or a convoy, it turned on its beacon, and the beacon could then be used by other aircraft or by surface vessels for homing. Operations against the Axis supply line to North Africa were carried on in this way. In the rendezvous of formations, a beacon in the lead aircraft permits assembly of the formation with ease.

*Close Control.*—As the strategy of the war shifted from the defensive to the offensive, the purpose of ground control of aircraft shifted away from interception of enemy bombers. Emphasis was placed on the control of large formations of our own fighters and bombers. The advent of high-resolution microwave ground radar made it possible to get a complete picture of the deployment of hundreds or thousands of aircraft, spread over an area of tens of thousands of square miles. Surveillance and control of this situation was facilitated by IFF beacons, and later by special microwave beacons designed for the special purpose of close control.

Used with air-surveillance radar, the beacons permitted the identification of individual airplanes or formations. With fire-control radar sets provided with suitable instrumentation (in particular, the SCR-584) control for tactical bombing with high accuracy was achieved.

## BOMBING SYSTEMS

**1·9. Oboe.**—One important use of airborne beacons was in a system called "Oboe" originally devised by TRE for precision bombing. In the Oboe system, two special interrogator-responsors were located at precisely known locations on the ground. Before the start of a mission, the distances between the ground stations and the target were computed with great accuracy. An aircraft carrying a beacon that would be interrogated by the ground stations was then flown toward a target. The system was so arranged that communication signals could be sent over the beacon system. These signals were used to instruct the pilot of the aircraft as to the course to fly, and also to instruct the bombardier as to when to drop his bombs. The Oboe system is illustrated in Fig. 1·9.

As used at night by the British, the beacons were installed in pathfinder aircraft—usually Mosquitoes—and these airplanes dropped flares on the target. The main force of bombers then bombed these flares by optical methods. When the U.S. Army's Ninth Air Force adopted the system, the pathfinder was the leader of a formation of Marauders bombing in daylight, and the formation dropped its bombs when the leader dropped his.

The tactical successes of Oboe were considerable; much of the bombing of the Ruhr in 1942 and 1943 was done with its aid. The story of these successes is told in detail elsewhere and we cannot pursue it further here. The system is noteworthy from a technical viewpoint in its utilization of the range extension provided by beacons, of the phenomenal precision of range measurement which is possible, and of the transmission of intelligence over the interrogation and response links.

*Courses Possible.*—In the Oboe system as it was used, the course flown by the aircraft is a circle. The center of the circle is at one of the ground stations, called the "cat" station; the circle passes through the correct point for releasing the bombs so that they will strike the target. This point is defined by its range from the other station, called the "mouse" station. Circular courses of this nature are called "cat-mouse" courses. They are not difficult for experienced pilots to fly, unless the radius is less than about 20 to 40 miles.

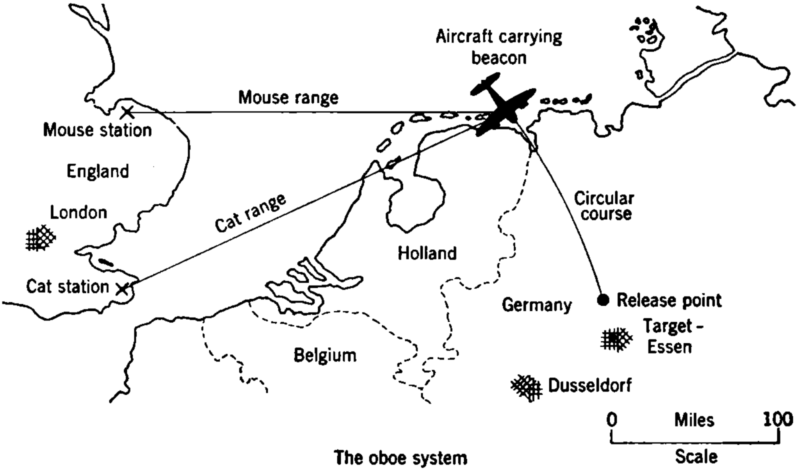


FIG. 1·9.—The Oboe system. An aircraft carrying a beacon flies at constant range from the cat station, until it reaches a predetermined range from the "mouse" station. The ranges are so determined as to bring the aircraft to the proper release point for the target chosen.

**1·10. H-systems.**—As everyone knows, the oboe is an ill wind that no one blows good.[1] The relevance of this bit of musical folklore to the Oboe bombing system lies in the fact that Oboe was operationally a difficult system and required virtuoso performances to achieve satisfactory results. The virtuosos were developed; the need for them was regrettable.

---

[1] See, for example, S. Fine, "Anatole of Paris," Decca C-91-4, 36583, Co 32301, U.S. Patent 1625705/1702564.

A solution for many of the difficulties encountered in the Oboe system is to be found in its converse, which is called the "*H*-system." In an *H*-system, the aircraft carries an interrogator and measures precise ranges to two ground beacons at accurately known locations. (See Fig. 1·10.)

*Courses Possible.*—In an *H*-system, the crew of the aircraft has control of the courses to be flown. With suitable instrumentation, a variety of choice is possible. Cat-mouse courses may be flown, as in Oboe. In addition, courses such that the difference of range to the two ground beacons is constant may readily be flown; such a course is a hyperbola.

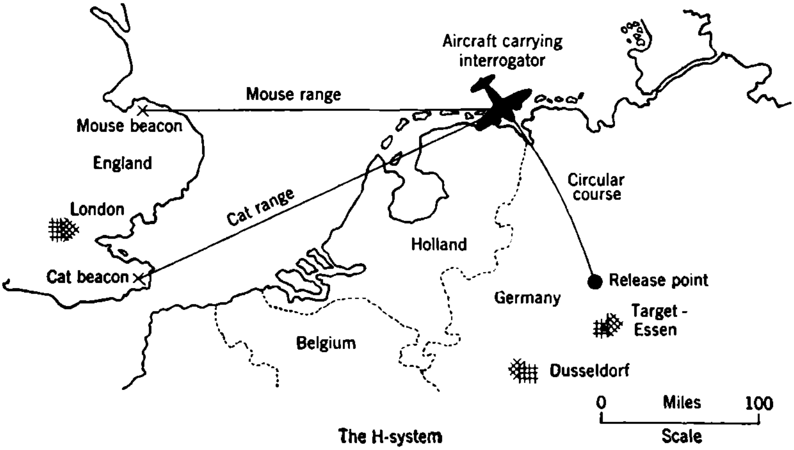


FIG. 1·10.—The *H*-system. This is the converse of the Oboe system (Fig. 1·10). The aircraft carries an interrogator, and the two ground stations are beacons.

With a suitable computer and instrumentation, arbitrary courses may be flown. Such computers were designed but not used in combat.

*Comparison of Oboe and H-systems.*—The major advantages of an *H*-system over an Oboe system are the following:

1. A very large number of aircraft can use the system at one time; in Oboe, one airplane is directed at a time by one pair of ground stations.[1]
2. Each plane can, if desired, attack a different target; many more targets are thus possible than with Oboe.
3. No communication to aircraft from ground stations, or between ground stations, is required; much less organizational effort is thus required to set up a mission.

---

[1] More complex Oboe systems which permitted simultaneous control of several planes were designed but not put into operation.

4. As noted above, a wider variety of choice of course to be flown is possible.
5. In the case of aircraft carrying bombing radars, little auxiliary equipment is required for beacon interrogation; the same aircraft, therefore, can be used for either radar or *H*-bombing. Oboe requires more specialized aircraft.

Several different *H*-systems were used in the war. We will examine these briefly.

*"Gee-H."*—This British system was made by altering ground stations of the Gee hyperbolic navigation system so that they operated as beacons. Additional equipment was added to the aircraft Gee receiver to convert it into a Gee-*H* beacon interrogator. Both the RAF and the Eighth Air Force used this system successfully during the last year of the war in Europe.

*Rebecca-H.*—By altering the airborne Rebecca indicator, provision was made for displaying two beacon responses simultaneously on an expanded delayed sweep. The standard Eureka beacons were used. Installed in photo-reconnaissance aircraft of the RAF, Rebecca-*H* was used to direct airplanes to locations at which reconnaissance photographs were required.

*Micro-H.*—The 3-cm bombing radars AN/APS-15 and AN/APQ-13 were designed with beacon interrogation provisions and with expanded delayed sweeps. They were thus readily adaptable to *H*-bombing. The Eighth Air Force used the AN/APS-15 (called "Mickey" by the crews) together with 3-cm beacons (sometimes called "Minnie") for *H*-bombing during the last nine months of the European war. This system was known more formally as Micro-*H*. Micro-*H* and Gee-*H* helped turn the tide toward victory in the battle of the Ardennes (the "Bulge").

*Shoran.*—Probably the first of all *H*-bombing systems to be designed, Shoran was also the best instrumented and the most precise. It was, however, the last to be produced and used. It consisted of an interrogator-responsor and beacons especially and solely designed for H-bombing. It was used with remarkable success by the Fifteenth Air Force in Italy and the Ninth Air Force in Germany, and was nearly ready for use against Japan when the war ended.

## HOW BEACONS ARE USED

We have reviewed briefly the major uses to which beacons were put during the war. Because beacons boast no peacetime history, an analysis of their wartime uses is the best guide to how they may be used in the future. Such an analysis may be made from at least two points of view. We can analyze beacon systems according to the functions they perform or according to the functions that need to be performed. A brief analysis from each of these points of view is given below.

## THE FUNCTIONS OF BEACONS

The data obtained from beacon replies differ somewhat from those obtained from a radar echo. Replies from beacons, for one thing, convey more information.

In the case of radar echo, the information obtained is entirely dependent upon the characteristics of the radar set and target. The range of the echo is measured with a precision characteristic of the timing circuits of the radar set. The azimuth is measured with a precision characteristic of the azimuth-determining equipment of the radar. Resolution in range depends upon the pulse duration; and resolution in azimuth depends on the antenna characteristics and the frequency used.

The beacon reply may be treated like an echo by the interrogator, as far as range and azimuth determination are concerned. If the beacon reply is coded, however, it conveys information about the beacon. It may tell, for example, where the beacon is; in this case coordinate information is transmitted. It may identify itself by conveying identifying information. Finally, if the beacon system is properly designed and used, it may communicate still other general information.

**1·11. Coordinate Fixing.**—Beacons can be used to mark and identify known positions on the ground so that an airplane can determine its position with reference to them. The combination of an airborne interrogator with ground-based beacons thus constitutes an additional aid to navigation, which is compared with other types in Secs. 1·18 to 1·22.

When a fix of only moderate accuracy (within about a mile or two) is needed, the determination of the range and bearing of one beacon suffices. For the utmost accuracy—determination of position to within 100 ft—distances to two beacons at known positions are measured simultaneously because range can readily be measured much more accurately than bearing. Systems of this latter type are useful in applications demanding great accuracy. These include bombing, mapping by airplane photography, in which high accuracy of fix of the camera is required, and similar applications. It is necessary, of course, that the beacon delay—the brief time required for it to reply—be made constant and that allowance be made for it.

Obviously, the converse application can be made. The location of a beacon-carrying aircraft can be determined accurately by one or more ground stations at known positions. As we have seen, some systems of precision bombing employed in the war used beacons in this way. In principle, the beacon was superfluous because the radar echoes of the aircraft might have sufficed. In practice, however, the beacons ensured reliable operation at long range and positive identification of the aircraft.

Shore beacons may be used with shipborne interrogators as lighthouses are used. Observation of a single beacon determines a fix. The ship gets not only the bearing, as it does from a lighthouse, but also the range. Shipborne interrogator-shore beacon systems should be especially useful under conditions of poor visibility. Beacons have not been widely used in this way so far, but the introduction of such a system is now contemplated by the U.S. Coast Guard.

The coordinates of the interrogator may be determined whenever a beacon is located at a known position and a fix is taken on it, and the coordinates of a beacon may be determined when a fix is taken on it by a stationary interrogator of known position.

When neither the interrogator nor the beacon is at a known location, relative coordinates are still of value for homing. Thus, a naval aircraft may home on an aircraft carrier without knowing its precise location and, indeed, while its location is changing. Similarly, aircraft may home on a ground beacon without caring where it is, as in Rebecca-Eureka.

Since a series of fixes determines a course, it is possible to set a course and navigate it by means of beacons. It is easy to remain within a mile of such a predetermined course, and even within a quarter of a mile without too much trouble. Greater precision requires some special instrumentation, such as an expanded sweep; with the aid of such instrumentation, however, a ship or airplane can remain within 100 ft of a predetermined course. By such methods, it is feasible to fly an aircraft by instruments to an airport, to let down through an overcast, and to emerge from it at an altitude of 100 ft, directly over the end of the runway.

**1·12. Identification.**—An airplane or ship equipped with a beacon of particular characteristics is thereby marked and identified to some extent, as we saw in previous discussion of IFF, the identification system used most widely during the war. Because aircraft were often detected by radar long before they could be seen, the problem was one of classifying the radar-reply pips as belonging either to friendly or to hostile aircraft. The radar sets were of many different types, operating over a wide range of frequencies; each one, therefore, had to be provided with a separate IFF interrogator-responsor working at one of the frequencies chosen for the IFF system. The interrogator was often synchronized with the radar set so that the interrogating pulses went out together, and the radar echoes and IFF replies arrived together and could be displayed on the same indicator tube to give proper correlation in range. In order to hinder the enemy in using captured transponders for deception, IFF beacons were coded in various ways. A thorough presentation of the special problems of IFF would require another volume. They will not be treated further in this book, partly because many of them do not arise in peacetime uses of radar or beacons, and partly because IFF has been considered, oddly enough, as particularly sacred from the standpoint of military security.

There is a peacetime problem of identification, however, that is closely related to that of IFF. It is the problem of identifying any particular one of many radar echoes in the neighborhood of busy airports when airplane traffic is being controlled through radar surveillance. A beacon with enough complication of coding to give positive individual identification of every aircraft seems to be needed. Fortunately, the problem is simpler than the IFF problem in one respect: It is not necessary that all aircraft have their identifying beacons turned on at nearly all times.

The identifying reply codes of fixed navigational beacons are assigned by a central authority and published for the information of those who use them. An aircraft navigator within range of such a beacon identifies the beacon by means of its code, and thus can locate the aircraft with respect to the beacon. The code of a shipboard beacon identifies the ship carrying it.

**1·13. Communication.**—By its very nature the beacon system involves the use of two nearly independent send-receive systems, as do all two-way communication systems. Furthermore, it is rather wasteful of its portion of the electromagnetic spectrum in that the r-f bandwidth required for satisfactory transmission of the simple pulses would permit the superposition of intelligence-conveying modulation. The various kinds of reply codes mentioned in Sec. 1·2 can be considered as communication of a rudimentary sort. So can interrogation codes that ensure that the beacons will reply only when properly interrogated. Coding of both kinds and several more elaborate ways of using the beacon channels for communication and remote control are discussed at length in succeeding chapters (Chaps. 5 and 11).

## TYPES OF BEACON SYSTEMS

**1·14. Air-to-surface Systems.**—A method of classification of radar-beacon systems that brings out some points of interest is according to the location of the interrogator and of the beacon. Thus, an air-to-surface system is one in which an airborne interrogator works with a beacon located on the surface of the earth.

The beacon in such a system may be used as a reference point for air navigation, as we have already seen. If so, it usually has a receiver of good sensitivity and a powerful transmitter so that it may be interrogated and received by all aircraft out to horizon range. Beacons of this kind have been installed on shipboard to enable aircraft to home on vessels, and other shipborne beacons have been used for identifying ships.

Light portable beacons with short ranges were used in the war to give the locations of isolated forces such as paratroop pathfinders, guerilla units, and secret agents.

Air-to-surface systems, then, are used mainly for navigation. When this navigation is of high precision, the system may be used for special purposes. One of these is bombing. The use of precision navigation as exemplified by bombing may be described in more general terms as the delivery of cargo at a designated location. The procedure is simplest when the location is designated by placing a beacon there, but it is still readily feasible when the beacons are elsewhere.

Another purpose of precision air navigation may be the making of a map. Extensive development of Shoran for the purpose of aerial photography was undertaken, and it is expected that excellent maps will be prepared from pictures taken in this way.

One other application is for air-sea rescue. Lightweight beacons were provided in life rafts for this purpose; in peacetime it appears that all lifeboats should be so equipped.

**1·15. Surface-to-air Systems.**—The principal use of surface-to-air systems is to aid in the surveillance and control of large numbers of aircraft, either from a fixed station on land or from a ship. This was one of the most important uses of radar during the war. The system was found to work better when the aircraft carried identifying beacons. A similar system including beacons might aid the control of civilian air traffic, but there would be obvious problems of installation, reliable maintenance, and correct operation of beacons in aircraft operated by numerous companies. Unsolved problems of controlling air traffic, however, are already limiting the operation of civilian airlines. If a radar-control system is to be used—and several such systems have already been proposed—then beacons will undoubtedly have their place in it.

Surface-to-air systems can be used for precision navigation of aircraft as are the air-to-surface systems described in Sec. 1·14. However, it is possible to do very accurate navigation with only one ground station when the control is exercised from the ground. This is because a ground station can obtain accurate azimuth information by means of conical scanning or other methods of comparing signals that are used for fire-control radar. Where only one ground station can be used conveniently, a system of this type is advantageous. The error is not independent of range, however, as is the case with triangulation systems that measure range only. The error in location introduced by an error in azimuth increases with range; the precision at very long ranges is not so good as with a two-station system. The error is about ¼ mile at 100 miles with an SCR-584.

Certain surface-to-air systems have been supplemented by automatic plotting boards that make a permanent and accurate record of the aircraft course. In addition, they can be used to direct aircraft on arbitrary courses with great precision and would be suitable for control of aircraft near airports.

**1·16. Air-to-air Systems.**—The principal use of air-to-air beacon systems during the war was for rendezvous of aircraft. When, for example, a submarine had been sighted, an airplane could patrol the adjacent region until it could be relieved or assisted. If it carried a beacon, the airplane coming to take over the patrol could home on the beacon and make contact with ease.

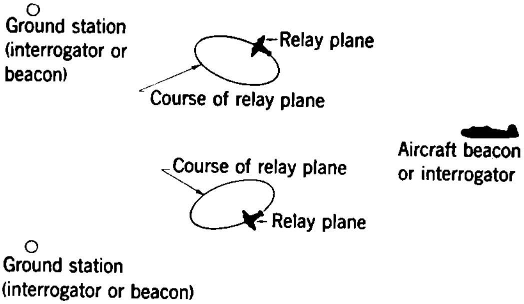


FIG. 1·11.—Beacon relay system. Interrogations and responses are relayed between ground stations and aircraft. The range of the system may be as great as 700 miles.

It seems unlikely that a peacetime need for a system of this sort will arise. One possible use, however, is for collision warning. Relatively simple equipment would make it possible for one airplane to have warning of the proximity of another beacon-carrying airplane. The same apparatus might function both as beacon and as interrogator-responsor. In this application however, the difficult problems of achieving universal use of the equipment again would be encountered.

*Relay Beacons.*—The maximum range of beacon systems is limited by the radio horizon. When one end of the system is on the surface of the earth, the range is determined by the altitude of the aircraft at the other end. At 30,000 ft the horizon range is about 250 miles. Air-to-air systems have the longest horizon ranges; 500 miles may be obtained. This property is of interest for relay systems. In such systems the effective range of an air-to-surface or surface-to-air system may be tripled, if the interrogation and response are relayed by an aircraft (see Fig. 1·11). In order to preserve the accuracy of ranges measured in such a system, either the courses of the relay aircraft must be restricted, or computers that make suitable corrections must be used.

As an alternative, an air-to-air beacon system may be used. In such a system the range of an aircraft from two other aircraft is determined. To convert such measurements into ground coordinates, the position of the two reference aircraft must be known at all times. These positions may be determined by an auxiliary radar or beacon system. Such a system is illustrated in Fig. 1·12.

A system of this type is of interest because it makes possible precision navigation anywhere within approximately 500 miles of identifiable radar landmarks. It does this at the expense of considerable complication, both of equipment and operational procedure.

During the war several relay systems were projected. An Oboe relay system, with an airborne relay, was actually used successfully by the RAF for at least one bombing operation in 1943.

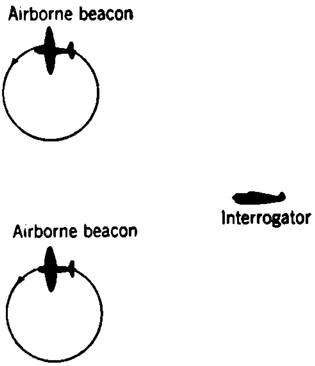


FIG. 1·12.—Airborne-beacon navigation system. The responses of the airborne beacons must convey information as to their instantaneous locations.

**1·17. Surface-to-surface Systems.** *Ground-to-ground Systems.*—Beacons have been used in surveying. With easily portable beacons and special interrogators, it is simple to get accurate measurements of the distances between widely separated points in fairly rough country. In this application it is necessary, however, to have a nearly clear line of sight from the interrogator to the beacon. At some lower frequencies, the beacon will be interrogated and will give a reply, even though there are sizable obstacles like hills in the direct line. The shortest distance of travel of the radio waves, however, is not the true distance because they have to travel around the obstacles.

Ground-to-ground surveying systems are usually designed to give measurements of range only, with precisions of 5 to 10 yds. They are usually short-range systems because of line-of-sight limitations. The equipment may be so designed that it can be used either as an interrogator-responsor or as a beacon by throwing a switch, or as both in rapid alternation with automatic switching between the two functions.

*Ship-to-shore Systems.*—The methods of precision navigation described above for aircraft may be applied to ship navigation. No such applications have yet been made, but they have been considered. For ship navigation off shore and in rivers, lakes, and harbors, systems can readily be designed which are adequate to permit navigation in well-defined lanes. In fact, plotting-table techniques can be used to permit continuous tracing of the position of the ship on a chart; they can even be extended to make possible the automatic steering of a ship along a predetermined course. Life on the Mississippi can now be made simpler for the pilot, even though somewhat less picturesque.

## COMPARISON OF BEACONS WITH OTHER NAVIGATIONAL AIDS

**1·18. Beacons and Ordinary Navigation.**—The use of beacons, especially with a radar indicator that gives a maplike presentation (plan position indicator, PPI) is in some ways merely an extension of ordinary nonradio navigation. Because beacons can be seen at much greater ranges than most landmarks can be seen visually, they give a much larger area of certainty of fix. An airplane with any standard airborne radar set can now fly the whole seaboard of the United States at an elevation of 5000 ft or more without ever being out of touch with one or more beacons.

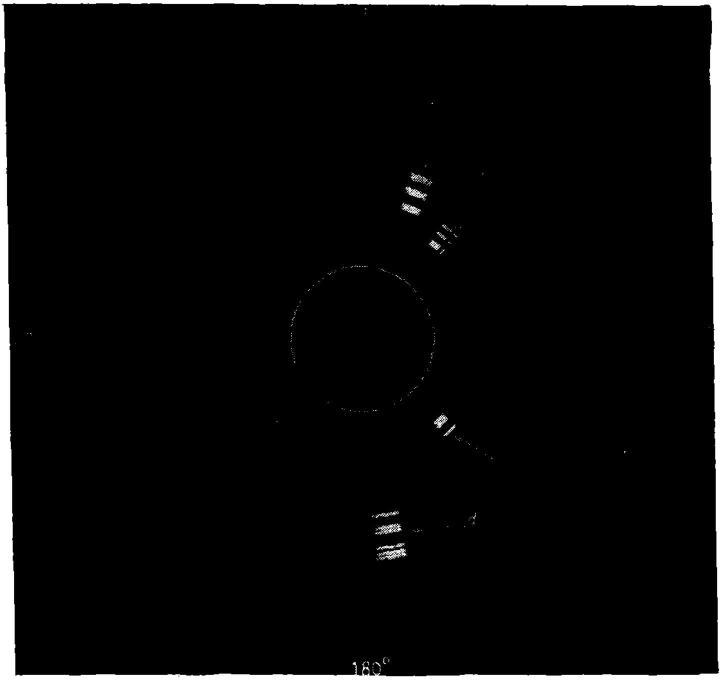


Fig. 1·13.—Four 3-cm beacons seen simultaneously on the PPI of an AN/APS-10 radar. Range circles are 20 miles apart. The beacons are as follows: (*a*) Code 2-2-2; Deer Island, Boston, Mass. (*b*) Code 2-1-1; South Weymouth, Mass. (*c*) Code 2-1; Quonset, R.I. (*d*) Code 1-2-3; Fisher's Island, N.Y. From a compass reading and the range and azimuth data obtained from any one of these beacons, it may be deduced that the aircraft is 2 miles north of Woonsocket, R.I., and headed toward Lexington, Mass.

This means that drift may be measured fairly accurately and checked frequently so that navigation becomes a nearly continuous process rather than a repeated checking of dead-reckoning predictions. Figure 1·13 shows a PPI photograph taken with an AN/APS-10 3-cm radar. Four 3-cm beacons are seen, at ranges between 30 and 50 miles. Each of these is identified by its code; any one of them would fix the position of the aircraft.

**1·19. Beacons and Radar Pilotage.**—Ordinary radar without beacons makes possible a similar type of pilotage if the terrain being flown over has distinctive configurations that do not present any ambiguities that hinder recognition. This fortunate condition, however, is not realized as often as is desirable; the addition of beacons to radar equipment gives much greater certainty of fix.

The inclusion of beacon-interrogation provisions in an airborne or ship radar set is a simple matter and is generally inexpensive. The augmentation of navigational facilities thus afforded is likely to be an excellent investment. Few American airborne radar sets were built during the war without such beacon-interrogating provisions. In these few cases, the special character of the radar sets made beacon operation impossible or impractical. (An example is the AN/APQ-7, which has beacon-interrogating provisions but is sometimes unable to use them because of the peculiar "squint" of the linear array antenna used.)

**1·20. Beacons, Radio Ranges, and Radio Compasses.**—In some ways the radio range is the simplest type of radio aid to navigation. The equipment carried in the airplane is simple and so is its use by the pilot. The information that it gives, however, is incomplete because it is merely a line of position. For adequate information, therefore, it has to be supplemented by other aids. Furthermore, since it gives only one line of flight, radio-range navigation does not permit any flexibility in choice of course.

The radio compass can indicate only the bearing of the transmitter being received. When two or more such bearings can be determined in rapid succession, a fairly accurate fix can be had, but this is not always possible. A fix is also obtained from one bearing taken while the airplane is flying a radio-range course. Both of these procedures involve some work with charts or computation.

An air-to-ground beacon-navigation system involves the use of more complicated apparatus in the airplane, but, in compensation, every reading of the range and bearing of an identifiable beacon gives a fix. When a PPI or type B indicator is used, the information is presented in a particularly simple and readily assimilable form. Furthermore, complete flexibility in choice of course is possible. Techniques have been worked out for navigation of an arbitrary course by the aid of a beacon. They are described in *Radar Aids to Navigation*, Vol. 2, Sec. 3·8, Radiation Laboratory Series.

**1·21. Beacons and Loran.**—In the Loran system and the Gee system, a moderately simple receiver-indicator is used in the plane. A reading of the difference of the time of arrival of pulses from two synchronized stations gives a location on a hyperbolic line of position. A second reading made on pulses coming from a second pair of stations gives another hyperbolic line of position. The intersection of the two lines gives the fix. The airplane must be equipped, in addition, with suitable charts giving Loran coordinates. This system permits navigation over long stretches of water, far out of the range of any possible beacons. Beacon-interrogating equipment is more complicated in design but simpler in use.

**1·22. Summary.**—The major characteristics of radar beacons, with an analysis of their uses, are summarized below.

1. Radar beacons are useful at ranges that may be limited only by the horizon. For air-to-surface or surface-to-air systems, such ranges depend on the aircraft altitude and are about 250 miles for aircraft at 30,000 ft. For surface-to-surface systems, ranges are, in general, considerably smaller.
2. At the frequencies used for beacons, no interference by atmospheric disturbances is encountered.
3. Beacons permit navigation of aircraft along arbitrarily chosen courses. A single observation on a beacon provides a fix. The accuracy of fixes taken with beacons may be as precise as desired, errors of less than 100 ft being readily possible.
4. Beacons permit ship navigation, within horizon range of land, with similar precision.
5. Long-range navigation very far from land is not readily possible at the frequencies used by beacons.
6. Beacons are excellent for homing. They may be so used at airports or for the delivery of airborne cargo.
7. Aircraft, flying predetermined courses, can use radio aids requiring less airborne equipment than is needed for interrogating beacons. Aircraft that do not adhere to rigidly predetermined routes may navigate arbitrary courses with ease by beacon methods.
8. Beacons installed in aircraft facilitate ground control by radar. Beacon responses can be displayed without interference from ground, sea, or cloud echoes, and convey much more information than echoes. They may be used for identification.
9. With some limitations, beacon systems can be used for communication.

## CHAPTER 2

# THE REQUIREMENTS OF SYSTEMS USING BEACONS: RANGE CONSIDERATIONS

BY H. H. BAILEY AND A. ROBERTS

In this chapter and the four following chapters the principles upon which the employment of radar beacons in pulsed systems is based will be discussed. These principles have been painfully evolved during the war and it may well be that they are not as yet completely understood, since the beacon art is still young. However, some fundamentals are clear; they will now be set forth. Other points are controversial; an attempt will be made to define them, even if they cannot be settled.

**2·1. Statement of Requirements.**[1]—A system that is to use beacons is projected in response to a need or purpose, generally expressed in the form of a statement of requirements for the system. Explicit statement of the requirements is always desirable, and, in fact, is usually essential for proper design of the system.

A statement of requirements for a system using beacons should specify the following:

1. Other uses of the interrogator, if any.
2. The reliable range and the degree of reliability desired.
3. The minimum range requirement, if any, and the coverage needed.
4. The expected density of beacons and interrogators.
5. The limitations of size, weight, power, etc.
6. The method of using the information obtained and the speed with which it must be made available.
7. The type of position information wanted and the precision required.
8. The amount of attention or maintenance available at the beacon.
9. The amount of information to be conveyed from beacon to interrogator.
10. The amount of interference to be expected.
11. The need, if any, for transmission of data (for example, control signals) over the interrogation and response links.

Succeeding sections of this chapter will discuss these requirements in detail.

---

[1] Secs. 2·1 to 2·3 by A. Roberts.

**2·2. First Principles of the Beacon Art.**—The most fundamental of all principles governing the design of beacons is this: *A beacon cannot properly be designed by itself, in isolation or in vacuo; the entire interrogating and reply system must be considered as a single unit.*

Self-evident as it may appear, this principle has been more often violated than observed. The exigencies of war have often resulted in systems that a conscientious engineer must abhor for their disregard of the simplest principles of efficiency or economy. Such systems have been made to work; often they have even worked satisfactorily in the field. But in any future design, to be governed by ordinary engineering and economic practices, it is essential to start from the standpoint that a beacon system has two ends that must be designed together for optimum efficiency, economy, and utility.

Beacons can be designed or modified to be used with interrogating equipment already in widespread use; likewise, interrogating equipment can be designed or modified to be used with beacons already in widespread use. Restrictions are likely to be encountered, however, which make the task difficult and the result only partially satisfactory.

One of the first questions to ask before designing a system using beacons is this: Will the interrogating equipment be used for radar echoes as well as for beacon responses?

If the interrogating equipment is to be used for radar echoes as well as for beacon responses there are two separate cases to consider: that in which the radar function is primary and the beacon-interrogation function secondary, and that in which the beacon function is primary and the radar function secondary. In the first case, the considerations determining radar performance will, then, in many particulars take precedence over beacon requirements (the choice of frequency, for instance). Since beacon requirements can be fulfilled more flexibly than radar requirements, a suitable beacon system can be designed for almost any contemplated radar system, after due consideration of requirements, cost, and probable performance. Beacon facilities cannot always be added, however, when the radar design has been fixed without regard for beacon performance.

If the radar function is secondary to the beacon function of a system, an interrogator-responsor can be so designed as to be useful as a radar, but its usefulness is likely to be limited. It may, for example, be restricted to collision warning.

If the interrogator is to be used solely with beacons, beacon requirements only will need to be considered. This fact, however, instead of simplifying the problem may make it more difficult since a large number possibilities is opened up.

It may happen that the beacon function is the more important, but a subsidiary radar function would be desirable if possible. In this case several alternate designs should be considered and a decision arrived at after consideration of the various alternatives.

Two lesser principles govern the design of systems using beacons. One is that *the use of microwaves for systems employing nondirectional or only slightly directional antennas at both ends is uneconomical.* It will be seen later that the power requirement for such systems increases as the inverse square of the wavelength.

The second principle is that *for systems using nondirectional antennas at one end and fixed antenna apertures at the other end, the power requirements are to a first approximation independent of wavelength.* The beamwidth of the fixed aperture antenna will, of course, decrease directly as the wavelength.

**2·3. Characteristics of Beacon Systems.**—The characteristics of a good beacon system designed for cathode-ray-tube display are

1. Accurate range (as accurate as desired).
2. Accurate azimuth (as accurate as desired).
3. Presentation of many beacons simultaneously without confusion.
4. Narrow azimuth arcs, good azimuth resolution.
5. Good over-all resolution, hence good identification.
6. Complete coverage, even at low angles of elevation.
7. Map type of presentation.
8. Simple correlation of echo and beacon responses.
9. Adequate traffic capacity for both interrogation and response.
10. Fast reading of beacon codes.
11. Ability to space beacons closely.

Many of these characteristics are more readily attained with microwave beacons than with lower-frequency systems.

*Azimuth Requirements and Frequency Choice.*—It is rarely desirable to use microwaves for systems used exclusively for beacon responses. If the interrogator and beacon both use a nondirectional antenna, microwaves are, as has been stated, uneconomical. If the interrogator uses a directional beam, a suitable interrogator-responsor has all the properties of a radar set and for maximum economy ought to be used as such in addition to being used for interrogation of beacons. This is true since even low (8-kw) pulse powers are useful for microwave radar; lower powers do not give proportionate reduction in size, weight, or power consumption.

It may, of course, happen that requirements for extremely accurate azimuth information will dictate the use of microwaves in a system intended only for beacon use. Thus microwaves might well be used in a ground interrogator for automatic tracking even if no radar use were contemplated. Such a set would nevertheless be useful as a radar set. It may also be necessary to use microwaves when spectrum space is not available at lower frequencies.

If fixes on a beacon are to be taken by range and azimuth, the superior directional properties of microwaves suggest their use, especially if any considerable degree of precision is required (about $\pm 1°$ in azimuth or better). If fixes are to be taken by range measurements only, using two or more beacons and triangulation methods, any frequency above about 30 Mc/sec is useful.

Good range and azimuth information can be obtained at lower frequencies as well as at microwave frequencies. Azimuth accuracy of plus or minus a few degrees can be obtained by lobe-switching techniques. The chief characteristic difference between systems that use lobe switching for azimuth accuracy and systems that obtain the same accuracy by using narrow beams is the much greater traffic capacity of the latter.

*Identification and Frequency Choice.*—For purposes of identification, microwaves are far superior to ultrahigh frequencies. Only frequencies that permit PPI presentation are capable of providing identification in any but the simplest situations.

The problem of instantaneous identification of any one beacon from thousands of others is still unsolved. It is certainly soluble, however; it is even possible that a solution will be attained within the next several years. The problem is essentially one of display. It is necessary that numerous coded replies be observable simultaneously; thus reply codes must not take much range on the PPI and must, at the same time, convey the desired information in a form readily observable by the eye.

## RANGE REQUIREMENTS

BY H. H. BAILEY

**2·4. Maximum Reliable Range.**—Range, of course, refers to the distance between the interrogating equipment and the beacon. There are both upper and lower limits to the range at which a system can function properly. A discussion of minimum useful range will be given later.

More important and much more fundamental is the maximum range, if this phrase is properly interpreted. The greatest possible range obtainable with a system has little practical significance, since it can be affected markedly by such things as performance of individual sets, temporary atmospheric effects, and the skill of the operator.

A far more useful concept is that of maximum reliable range, or simply the *reliable range.* This is the greatest range that can ordinarily be counted upon under practical conditions with average personnel operating production equipment. A precise definition of reliable range, from its very nature, cannot be given. However, excellent correlations between experiment and theory have been obtained in the case of scanning radars by considering as reliable *that range at which a good radar operator can see the beacon on three-fourths of the sweeps.*

The reliable range of a complete system is a property that, more than any other, depends inherently upon the characteristics of both sets in the system. Indeed, it is affected by a great many factors. Futhermore, a given range can be obtained by many different designs. For these reasons, it is not possible to treat the desired range as a design parameter and then attempt to show how best to design a system having that range. Rather, considerations of size, weight, cost, availability of components, and other required functions, as applied to the two parts of the system separately, must be adjusted to fit the overriding requirement of range. Accordingly, the point of view adopted here will be to show how the reliable range to be expected can be computed when all the detailed characteristics of the interrogator and the beacon are known.

**2·5. The Range Equation.**—All of the systems under consideration make use of two distinct propagation links: the interrogation link and the response link. It is clear that the controlling equations will be those for one-way transmission applied to the two links separately (as differentiated from the round-trip formulas applicable to radar echoes). Obviously, the reliable range of the system will be the lesser of the ranges computed for the two links. The transmission equations for free-space conditions may be derived as shown in the following paragraphs.

Consider an isotropic radiator emitting a total power $P_T$ watts. The power radiated in any direction is $P_T/4\pi$ watts per steradian. If the radiator is replaced by an antenna with a gain (over isotropic) of $g_T$ in the desired direction, the power radiated in that direction is $P_Tg_T/4\pi$. If a receiving antenna with an effective cross section $a$ is placed at a distance $R$ from the transmitter, it subtends a solid angle $a/R^2$ steradians, so that the power intercepted by such an antenna is $P_Tg_Ta/4\pi R^2$ watts. In order to achieve symmetry and greater generality, $a$ is expressed in terms of the gain of the receiving antenna $g_R$.[1] The needed relationship, which is derived in *Microwave Antenna Theory and Design,* Vol. 12, Chap. 6, Radiation Laboratory Series, from a straightforward computation of the power pattern of the antenna, is that $g_R = 4\pi a/\lambda^2$. Thus the power received is

$$P_R = \frac{P_Tg_Tg_R\lambda^2}{(4\pi R)^2}.$$

It will be found convenient to use only the maximum gains $G_T$ and $G_R$ of the transmitting and receiving antennas respectively, and to apply a

---

[1] For a Hertzian half-wave dipole, the gain is $\frac{3}{2}$ and the equivalent area $3\lambda^2/8\pi$.

correction factor whenever the geometry of a situation is such that the full gain of an antenna is not utilized. It will also be appropriate for the present purposes to solve the above equation for $R^2$. The range equation then becomes

$$R^2 = \left(\frac{\lambda}{4\pi}\right)^2 \frac{P_T G_T G_R}{P_R}. \tag{1}$$

Considerable care must be exercised to combine or modify suitably the quantities that are directly measured or computed before inserting them into this equation. It is, therefore, necessary to rewrite the equation, as applied to each link separately, in terms of readily measurable quantities, including all losses. In order to do this, a rather complete system of notation must be established:

Let $R$ = reliable range as defined above, in any units.
$P$ = power, in watts.
$G$ = maximum effective gain of an antenna over isotropic.
$L$ = transmission line loss, in decibels.
$M$ = reduction from maximum gain to utilized gain, in decibels.
$A$ = atmospheric absorption, in decibels.
$\lambda$ = wavelength of radiation used, in same units as $R$.
$K$ = loss factor [see Eqs. (2) and (3) below].
$i$ subscript refers to interrogation link.
$r$ subscript refers to response link.
$T$ subscript refers to transmitting components.
$R$ subscript refers to receiving components.
Primed quantities refer to the beacon.
Unprimed quantities refer to the interrogator.

Equation (1) now gives rise to two equations, as follows:

$$\left.\begin{aligned} R_i^2 &= (\lambda_i/4\pi)^2 (P_T G_T G'_R / P'_R K_i) \\ \text{where} \quad K_i &= \text{antilog}_{10}\,(L_T + L'_R + M_T + M'_R + A_i)/10 \end{aligned}\right\} \tag{2}$$

and

$$\left.\begin{aligned} R_r^2 &= (\lambda_r/4\pi)^2 (P'_T G'_T G_R / P_R K_r) \\ \text{where} \quad K_r &= \text{antilog}_{10}\,(L'_T + L_R + M'_T + M_R + A_r)/10. \end{aligned}\right\} \tag{3}$$

Another useful formulation is the set of equations, Eqs. (4) and (5) below:

$$10 \log_{10} P'_R = 10 \log_{10} \frac{P_T G_T G'_R \lambda_i^2}{(4\pi R_i)^2} - (L_T + M_T + A_i + M'_R + L'_R). \tag{4}$$

This gives the power, in decibels below 1 watt, available at the input terminals of the beacon receiver.

$$10 \log_{10} P_R = 10 \log_{10} \frac{P'_T G'_T G_R \lambda_r^2}{(4\pi R_r)^2} - (L'_T + M'_T + A_r + M_R + L_R). \quad (5)$$

This gives the power, in decibels below 1 watt, available at the input terminals of the interrogator receiver.

*Interpretation of the Quantities in the Range Equation.*—The loss terms require discussion. The line losses, including duplexing losses if they exist, are directly measurable and cause no difficulty. $M'$ is often zero, since beacon antennas usually have their maximum gains close to the horizontal and so make use of the full gains at maximum range. $M$ is arbitrarily taken as 3 db so that signals will be considered reliable when they occur over an arc equal to the half-power beamwidth of the antenna. An experimental justification for this value of $M$ is given later (Sec. 2·9).

The absorption $A$ of the atmosphere is a function of wavelength, pressure, and humidity as well as range, and the rate of absorption may vary considerably along the path of the radiation (as when thunderstorms are present). Nevertheless it can usually be estimated with sufficient precision and in most cases is actually negligible.

The interpretation of the power terms also needs some clarification. The transmitted power refers to the pulse power as measured at the transmitter itself. The received power refers to the *effective* receiver sensitivity. In the beacon receiver, it has been found experimentally, as discussed in Sec. 8·3, that a signal must have an amplitude at least 4 db above rms noise[1] to produce consistent triggering of the succeeding circuits when the gain is set so that triggering on noise itself is negligible. The sensitivity of a beacon receiver is therefore quoted in terms of the reliable tripping level, and this quantity is $P'_R$ directly, by definition.

$P_R$ is the power arriving at the radar receiver which produces a reliable result. This is different, in general, from the advertised value of the receiver sensitivity; consequently the advertised value must be corrected. The incoming signal required, even by identical receivers, depends markedly on the final result that is to be achieved. Sets such as beacons themselves, in which the signals actuate electrical or mechanical detecting devices, respond to individual pulses and can be made to operate on signals comparable to noise peaks or even less if the noise firing itself is not harmful. On the other hand, in sets that must average or integrate the received signals for automatic tracking purposes, random triggering is quite harmful. Such sets may have to operate at signal-to-noise ratios of about 10 or even higher. The exact value is a function of the circuits involved and of the accuracy of tracking required. The widest application is to sets that provide a visual indication on some kind of a cathode-ray tube, and these will be discussed at greater length.

---

[1] This figure applies for a square-law receiver. For a superheterodyne, the value is 8 db.

## Display Losses

**2·6. Losses in CRT Displays.**—In sets using CRT displays, the desired final result is not achieved until an operator has noticed or turned his attention to the presentation of a signal. Thus, two very complicated sets of phenomena become involved—various characteristics of the CRT screen and the physiology and psychology of a human observer. This extremely difficult subject has been investigated in experiments that are described in *Threshold Signals*, Vol. 24, Chaps. 7, 8, and 9, Radiation Laboratory Series. Certain facts and relationships have emerged which are very useful in deriving the corrections that must be applied to measured values of receiver sensitivity before they can be used in Eq. (2).

*Minimum Discernible Signals.*—The strength of a signal that can just be discerned in noise varies inversely as the square root of the pulse-repetition frequency, other things being equal. This is true for both deflection-modulated and intensity-modulated CRT displays. Thus, if under certain conditions a signal-to-noise ratio of 1 is necessary at 80 cps, a signal at 400 cps such that the signal-to-noise ratio is $1/\sqrt{5}$ can still be seen, the gain in sensitivity being 3.5 db; 7 db of sensitivity are gained over the original if a rate of 2000 cps is used.

The same law holds even when the pips are not superimposed. That is, a signal that occurs on four times as many sweeps as a second signal can be detected in noise when it has half the strength of the second signal, even when the successive sweeps are displaced, as in B-scope or PPI presentations. It may be noted that the conditions for these two laws become identical in A-scope presentation. In fact, the two laws are but different aspects of a fundamental relationship that has been derived by G. E. Uhlenbeck[1] from purely statistical arguments on the comparison of a signal with the appropriate amount of noise.

*Scanning Losses.*—The importance of the second relationship lies in its application to scanning. Thus, a minimum detectable signal under scanning conditions must be greater than the corresponding signal under "searchlighting" conditions by a factor that is the square root of the ratio of the total angle scanned to the angle over which the signal is actually seen (nominally, the horizontal beamwidth). For example, consider a receiver with a certain sensitivity as measured with a signal generator under nonscanning conditions. Used with an antenna that has a beamwidth of 3° and scans through 150°, it will suffer an 8.5-db loss (i.e., $\sqrt{\frac{3}{150}}$) in sensitivity. This square-root scanning loss is actually subject to the restriction that the scanning period be less than about 6 to 10 sec. However, the transition from the above law to no loss at all

---

[1] *Threshold Signals*, Vol. 24, Chap. 7, Radiation Laboratory Series.

(for extremely slow scanning rates) sets in gradually and begins to be effective in reducing the scanning loss only on the large ground radars.

*Pulse Width.*—One additional correction, usually small, is for pulse width. The beacon signals may have in general a different, and usually a shorter, pulse width than those with which receiver sensitivity is ordinarily measured. The importance of this difference depends, of course, on the bandwidth of the receiver. The correction is most easily determined by a straightforward sensitivity measurement using signals with the actual beacon pulse width.

The experimental curves of minimum discernible signal strength vs. width of the indication on the screen (see Fig. 2·1) are found to be concave upward with a broad minimum at about 1 mm. This result, at first sight rather surprising, is almost independent of CRT focus or spot size and seems to be related to the resolving power of the eye. At signal widths less than 1 mm—that is, at slower sweep speeds—the familiar square-root law is approximately followed. For example, a 0.5-$\mu$sec pulse viewed with a 100-mile sweep on a 5-in. PPI tube is 0.025 mm wide, and the curve indicates that the effective sensitivity obtainable under these conditions is 4.8 db below the optimum and 4.5 db below that measured on a 4-mile sweep (signal 0.6 mm wide).

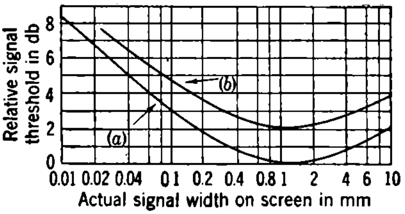


FIG. 2·1.—Relative intensity of minimum discernible signals as a function of signal width on the CRT screen. (a) A-scope; (b) B-scope or PPI.

*Response Losses.*—Even granted adequate power and sensitivity on the interrogation link, 100 per cent response by the beacon is not assured. This effect, unlike that for radar reflections in which there is a statistical distribution of intensity of the reflected signal, involves the phenomenon of missing responses. It is produced because all beacons have a "blanking gate" that produces a "dead" time. Under conditions of interrogation of a beacon by a number of interrogators, the per cent response $W$ can, in principle, be computed.[1] Then, theoretically, $P_R$ should be increased by the factor $1/W^{1/2}$, or the corresponding range decreased by $W^{1/4}$. Actually, beacons should be designed with sufficient traffic capacity for a given application (a discussion of the parameters controlling this is given in Chap. 6); thus, in practice, $W$ will seldom assume values low enough to cause a significant reduction in range. This effect will be neglected in the remainder of the present section.

*The Observer.*—Finally, there is the observer to consider. Semitrained personnel, who are usually depended upon in the use of specialized elec-

---

[1] See Chap. 6.

tronic equipment, differ greatly in their skill and motivation. Factors like ambient noise and ambient light have an important effect on the attentiveness required to achieve maximum performance. The supply of oxygen or lingering effects of the previous night's diversion can alter the degree of attention of which an operator is capable. These variable effects obviously cannot be evaluated quantitatively, but they do make it advisable to provide a margin of 10 db or so more power or sensitivity than would otherwise be indicated, when the additional cost, weight, power, time, and so forth, are not prohibitive.

*Summary.*—To ecapitulate, the value to be used in Eq. (2) for $P_R$—the *effective* sensitivity for producing a desired result "reliably"—must be obtained from the measured value of sensitivity by making appropriate allowances for probable conditions of use and the final result desired. In particular, if a visual indication is to be presented to an observer, corrections must be made according to known laws when the conditions under which the quoted sensitivity was measured and the actual conditions in the field differ in pulse width, scanning rate, or sweep speed. If possible allowances must be made for the differences between real observers and ideal observers.

Sometimes receivers are used for functions other than beacon reception, such as radar reception. Their sensitivities are often measured by maintenance personnel under almost optimum conditions (long pulse, high repetition rate, no scanning, fast sweep) so that optimistic values are quoted. The corrections are therefore usually necessary, and they may be sizable.

**2·7. Losses in Other Displays.**—Radar information, whether obtained from echoes or from beacons, is ordinarily displayed on CRT screens. However, this information can also be displayed in other ways. The most important of these are probably the use of meters or the use of motor-driven indicating mechanisms. Rows of lights are used on occasion; relays or other trigger devices also may be employed.

Such indicating devices are invariably operated by circuits that detect the received pulse in the presence of noise. They do this by comparing the signal plus noise to the noise alone. It is immediately clear that such circuits are free from most of the defects of the human observer using a cathode-ray tube, and are best compared to the expanded A-scope presentation because of their ability to ignore irrelevant noise. Accordingly, one expects the losses in such circuits to be low compared to CRT losses, which is actually the case. As an example, a pulse-width discriminator circuit that has been used in beacons will trip reliably when the signal is 4 to 5 db above noise. For other circuits the margin may be even less. This is to be compared with CRT display losses, which may exceed 15 db; signals must, therefore, be 10 db or more above noise to be reliable. When the general noise level is subject to short-term fluctuation, however, the human observer may do better because of his intelligent selection of the intervals of time to be taken as significant.

**2·8. Video Stretching.**—It has been seen that a presentation loss is incurred if the length of the signal on the CRT sweep is different from an optimum value of about 1 mm. Figure 2·1 shows the loss quantitatively as a function of signal length. In many systems, much of this loss can be avoided by the use of an expedient known as "video stretching."

If it were possible to lengthen the beacon reply pulse to 3 to 5 $\mu$sec, it is clear that on long sweeps much of the loss could be avoided. A 10-mile sweep on a 5-in. PPI is about 60 mm long and covers 120 $\mu$sec; a 0.5-$\mu$sec pulse is then 0.25 mm long. On a 100-mile sweep on the same tube it is only 0.025 mm long. The corresponding loss compared to a 1-mm signal is 5.5 db. If the pulse were 5 $\mu$sec long the loss would be 1.0 db; this is an improvement of 4.5 db.

Now the use of 5-$\mu$sec reply pulses rather than a 0.5-$\mu$sec pulse would increase the beacon duty ratio by a factor of 10; the decrease in traffic capacity would be intolerably great in cases in which the duty ratio is a limiting factor. For microwave beacons this is almost always true. The equivalent effect for strong signals can be achieved by so arranging the video circuits of the interrogator receiver that incoming pulses are

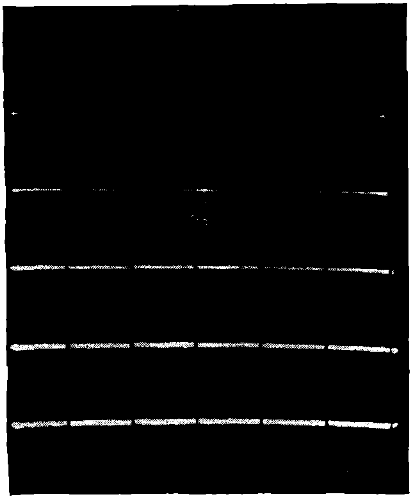


FIG. 2·2.—B-scope showing beacon signals without video stretching in the interrogator receiver. The range marks are 20 miles apart.

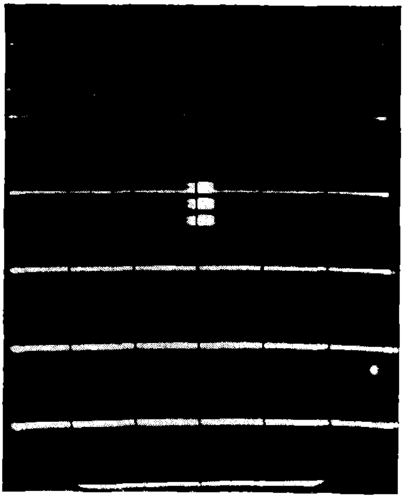


FIG. 2·3.—B-scope showing beacon signals with video stretching in the interrogator receiver. The range marks are 20 miles apart. The increased illumination of the screen is of most value for long sweeps. Note that range measurements of the leading edge of the signal are unimpaired.

"stretched." The stretching should be done in such a way that range accuracy is preserved; thus, the leading edge of the pulses should be unaltered. The rest of the pulse should be stretched out so that it appears like a 2- to 3-μsec pulse on the cathode-ray tube. Longer stretching interferes with range code legibility. Figures 2·2 and 2·3 show the effect of video stretching on code legibility.

It should be clearly understood that video stretching is effective in increasing the *reliable* range, not the *maximum* range. It can only eliminate or reduce the long sweep loss. It cannot change the maximum range because it does not affect the ratio of signal to noise. In systems working with signals very close to noise, video stretching will introduce no improvement; with scanning microwave radar sets and average radar operators, a very considerable improvement is introduced. Field experience confirms this.

In systems using automatic range-tracking, or any signal detector other than the cathode-ray tube, video stretching is unnecessary and undesirable.

## Applying the Range Equation

**2·9. Experimental Verification of Reliable Range.**—In order to design a certain lightweight beacon so that it would just meet and not exceed the requirements for performance, an extensive series of carefully controlled flight tests was conducted by the beacon group of the Radiation Laboratory. An airborne scanning 3-cm radar was used to interrogate an experimental 3-cm beacon; the effective power output and receiver sensitivity of the beacon could be adjusted at will over a fairly wide range of values. These experiments constitute an excellent test of the range equation and all the corrections that must be applied to measured values in the use of that equation. They have verified the theory completely within the experimental error. Other isolated cases in which all of the necessary data were available have also given similar checks, but this particular series of tests is sufficiently significant to warrant noting some of the results here.

The beacon contained a more powerful transmitter and a more sensitive receiver than necessary and was fitted with calibrated attenuators in both lines. Thus, two things could be accomplished: By removing the attenuation from one line, one link could be made very strong to ensure that the observations of range would be affected by only one factor at a time; by varying the other attenuator, a variable power output or receiver sensitivity could be simulated. Under a variety of conditions, an experienced radar operator in the aircraft made runs approaching the beacon and recorded, whenever possible, both a maximum and a "reliable" range on each run. The latter was, by definition, that range at which this particular operator could "see" the beacon on three-fourths of the scans past the beacon. His observations when testing the response link are indicated in Fig. 2·4. It should be noted that the maximum range data have a much greater spread than the reliable range figures—an additional reason for considering reliable ranges the more significant.

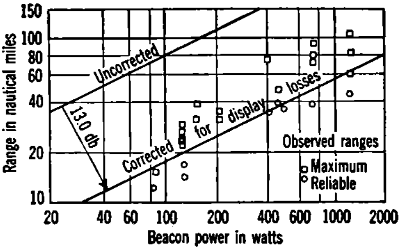


FIG. 2·4.—Experimental verification of the range equation, with display corrections. (For details of the corrections, see text.) The two lines are drawn with an inverse square slope. The upper line represents uncorrected predicted range. The lower line shows the predicted range with the calculated display correction of 13.0 db. Data are for the response link of a 3-cm airborne-radar–ground-beacon system.

The upper diagonal line, marked "uncorrected," represents the theoretical dependence of range on output power with none of the required display corrections made to Eq. (2). In order to simplify the argument, the attenuator readings are used to adjust $P_T$. System corrections include a measured loss in the beacon r-f transmission line up to the antenna of 0.5 db, a negligible radar r-f line loss, 3 db loss taken (arbitrarily) in $M$, atmospheric absorption neglected; hence there was 3.5-db loss. Interference nulls were minimized by flying high.

The lower line takes into account the CRT display corrections to receiver sensitivity which are computed as follows. The receiver sensitivity was measured by feeding in 1-$\mu$sec pulses at 500 cps and observing the indication on an A-scope with a 4-in., 15-mile sweep. The flight conditions, on the other hand, were 0.5-$\mu$sec pulses at 400 cps viewed on a B-scope using a 4-in., 100-mile sweep, with the radar's 5°-beam scanning through 150°. The corrections are then: pulse width 0.2 db (not measured in this case but computed from the known i-f bandwidth), repetition rate 0.5 db (from $\sqrt{500/400}$), sweep speed 4.9 db (from Fig. 2·1) for 0.27- and 0.04-mm signals (including the small difference between A-scope and PPI or B-scope presentations), and scanning 7.4 db (from 150°/5°), for a total of 13.0 db. Note that every one of the corrections is in the direction of reducing the effective sensitivity.

The correlation that Fig. 2·4 exhibits between the finally computed ranges, after all known corrections have been made, and the experimental observations based on an arbitrary three-fourths rule is better than could be hoped for. The best straight line (of proper inverse square slope) through the observed points is within 1 or 2 db of the theoretical curve, and that difference is certainly less than the probable error in the absolute values of the two r-f power measurements.

Similar tests made under conditions in which the limiting was in the interrogation link checked the theory equally well. For both links, the mean of the maximum ranges differs from the average reliable range by about 50 per cent, equivalent to 3 to 4 db. The value of 3 db which was adopted for $M$, the radar antenna pattern loss, under reasonable but rather arbitrary assumptions, can now be considered as justified by this experimental determination of the difference between maximum and reliable ranges.

**2·10. General Application to a System.**—After ranges have been computed by Eqs. (2) and (3) for the interrogation and response links respectively, the smaller of the two is, of course, adopted as the reliable range of the system. The link with the smaller range is known as the "limiting link."

If the interrogation link limits, the response is comparatively strong. This is advantageous, particularly for interrogators with a visual presentation, as has just been explained. It is expensive, however, from the beacon point of view, since transmitter and modulator power are expensive in terms of input power and hence in space and weight. If the response link limits, more skill is required of the interrogator operator in observing signals that are down in the noise.

Obviously, any appreciable unbalance between the two links is wasteful and should be avoided. The condition for balance is easily obtained from Eqs. (2) and (3), for, if the ranges are equal, the right-hand sides of those equations are equal to each other. In many practical cases, $\lambda_i = \lambda_r$, the antenna gains are the same on both links, and, if there is no large duplexer loss, even the $K$'s may be approximately equal. In such cases, the condition of balance reduces to the so-called "power product theorem,"

$$P_TP_R = P'_TP'_R. \tag{6}$$

Actually this relationship must be used with caution. There is not only the restriction that the $K$'s should be approximately equal, but also that for $P_R$ the corrected radar receiver sensitivity must be used. It can nevertheless be useful as a rough criterion in the early stages of design.

**2·11. Minimum Range.**—As the range between an interrogator and a beacon decreases, only the quantities $R$, $M$, and $M'$ in Eqs. (2) change. In general, two effects take place. At medium and fairly short ranges the vertical angular relationships, and hence $M$ and $M'$, change very little; since $R^2$ decreases manyfold, the strength of the received signal on both links increases by many decibels. At short ranges (a few miles) in systems in which one of the sets is airborne, the losses in the $M$'s, caused by operation down on the sides of the antenna patterns, become serious. They may even overcome the inverse square effect of the range to produce a "cone of silence" directly over or under the beacon.

The predominating effect of the inverse square law in raising the level of the received signal at decreased ranges has two serious consequences that must be considered in the design of a system. First, the beacon receiver must be capable of receiving very strong signals, 60 db or more above minimum, without being blocked or otherwise rendered insensitive for an appreciable period of time. This requirement cannot be avoided; the interrogator-transmitter power might be varied appreciably, but such variation would not always be desirable because there might be another beacon at maximum range in the same azimuth. Similarly, the beacon receiver should not have its gain or sensitivity varied for the analogous reason that there might be an interrogator at maximum range trying to get responses at the same time.

Second, signals may be received over a very wide arc, even 360°, because the $M$ factor of the interrogator antenna can decrease as much as $R^2$ decreases, and operation will still be maintained. This amounts to reducing or even removing the azimuth information just when, for some applications, it may be needed most of all. Various methods can be used to minimize this effect. The operator may simply reduce his receiver gain (thus increasing the required received power $P'_R$) or change the tilt of his antenna (thus decreasing $M$) when he wishes to concentrate on a near-by beacon. Either procedure, of course, sacrifices replies from more distant beacons; this may or may not be serious.

It is also possible to install on the receiver an automatic variable gain control, which markedly reduces the gain immediately following the transmission of each pulse and then allows the gain to recover in such a manner that near-by strong responses and weaker signals that arrive later from more distant beacons can all cause satisfactory indications out to maximum range. This procedure is relatively inexpensive and produces excellent results. It is called "time-varied gain," TVG, and is described at length in *Microwave Receivers*, Vol. 23, Chap. 9, Radiation Laboratory Series.

The "cone of silence" phenomenon is controlled entirely by the two antenna patterns and the range of tilt that are available. Both patterns ordinarily have very low gain in the vertical direction, since in both cases primary emphasis is placed on the horizontal direction or on moderate elevations or depressions. It is, however, possible to design a beacon antenna to overcome this effect (see Sec. 7·3).

# CHAPTER 3

# PROPAGATION AND COVERAGE CONSIDERATIONS

BY A. ROBERTS

## PROPAGATION EFFECTS

Interrogators and beacons do not operate in free space. It is now time to consider this fact seriously. Both interrogators and beacons are close to a nearly spherical earth and are immersed in an atmosphere of air that has pronounced and not always predictable effects on signals. The signals are reflected or scattered from the earth, refracted and absorbed by the atmosphere, reflected and refracted by boundaries between significantly different portions of the atmosphere, and diffracted by obstacles, including the horizon. Some of these effects will now be considered briefly.

**3·1. Coherent Interference.**—Whenever coherent waves from an emitter reach a remote point by two or more paths, interference occurs. This effect is well known for light waves, and is the basis of methods of measuring the wavelength of light.

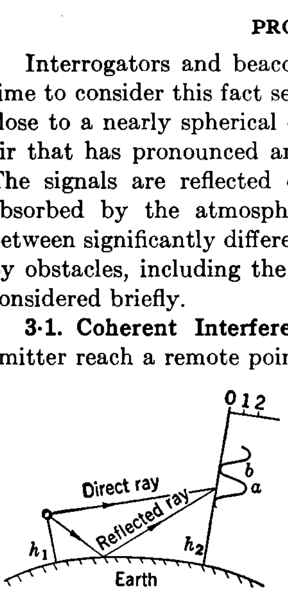


FIG. 3·1.—Coherent interference due to reflection from the surface of the earth (the Lloyd's mirror phenomenon). (*a*) Interference maximum: phase difference = $n\lambda$. (*b*) Interference minimum: phase difference = $(n - \frac{1}{2})\lambda$. The pattern propagation factor $|F|$ is plotted as a function of altitude at a fixed range.

The signal from a transmitter can reach a receiver by two paths when the path between them is unobstructed and part of the transmitted signal reaches the receiver after reflection by the earth. When this effect takes place, the resultant signal at the receiver may have a value widely different from what it would have been had only one signal arrived. The reason is that the two signals are *coherent*, that is, related in phase. The amplitudes of the two signals must therefore be added vectorially; the resulting amplitude can be either greater or smaller than the free-space value, as shown in Fig. 3·1. The variation of amplitude gives rise to a series of "lobes," each of which corresponds to a maximum in the interference pattern.

In the simple case of a flat reflecting surface and radiation of equal intensities on both paths, the effect can be described by the relation

$$|F| = |1 + \rho e^{j\alpha}| = \sqrt{1 + \rho^2 + 2\rho \cos \alpha}, \tag{1}$$

where $F$ = "pattern propagation factor," the ratio of the resultant electric-field amplitude at the point in question to the value it would have under free-space conditions,

$\rho$ = magnitude of the reflection coefficient of the reflecting surface (for the particular angle of incidence, the frequency, and the polarization of the incident radiation), and

$\alpha$ = phase difference between direct and reflected rays.

The phase difference $\alpha$ is composed of two parts—the phase change on reflection, $\phi$, and that due to the path difference $\Delta R$ between the direct and reflected rays, which is $2\pi\Delta R/\lambda$. Thus $|F|$ varies between $1 + \rho$ and $1 - \rho$, as we can see from Fig. 3·2.

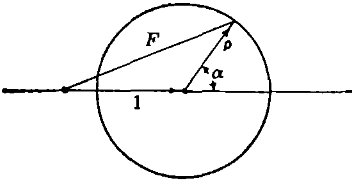


FIG. 3·2.—Graphical representation of $F$ as a function of $\rho$ and $\alpha$. $|F| = |1 + \rho e^{j\alpha}|$.

Clearly, the effect is of greatest consequence in the simple case in which there is only one reflected ray, and $\rho$ is nearly equal to 1; $F$ accordingly varies between its limits of 0 and 2. When $\rho$ is much less than 1, $|F|$ varies only slightly about a mean value of unity.

Since the power reaching the receiver depends on $F^2$, it can be expressed as

$$P_R = P_T \frac{G_T G_R \lambda^2}{(4\pi)^2 R^2} F^2, \tag{2}$$

and may thus vary between zero and four times the free-space value. In terms of decibels, the signal may be as much as 6 db more intense, or any number of decibels less intense, than the free-space value.

We have seen that the important term that determines just what the intensity at a particular point will be is $\alpha$, the phase difference between the two paths, and that

$$\alpha = \phi + \frac{2\pi\Delta R}{\lambda}. \tag{3}$$

For heights of receiver and transmitter $h_1$ and $h_2$, the path difference $\Delta R$ (neglecting terms in $(h/R)^4$ and higher powers) is, assuming a flat earth,

$$\Delta R = \frac{2h_1h_2}{R}. \tag{4}$$

Thus

$$\alpha = \phi + \frac{4\pi h_1 h_2}{R\lambda} \tag{5}$$

and

$$|F| = |1 + \rho e^{j\left(\phi + \frac{4\pi h_1 h_2}{R}\right)}|. \tag{6}$$

From Eqs. (2) and (6) the value of $P_R$ can now be calculated at any point if the following quantities are known in addition to system parameters:

1. The phase change on reflection $\phi$. This depends on the nature of the reflecting medium, the angle of incidence, the frequency, and the polarization of the incident radiation.[1]
2. The magnitude of the reflection coefficient $\rho$. This depends, both theoretically and experimentally, on the grazing angle of incidence, the polarization of the incident radiation, the degree of irregularity of the reflecting surface, and the electrical properties of that surface. The degree of irregularity, which determines the resemblance of the surface to a mirror, depends on the wavelength of the radiation. A surface that is smooth for radiation of 1-m wavelength may be rough for 10-cm radiation. This is of especial importance in land reflections.

When the path difference $\Delta R$ is small, the angle $2\pi h_1 h_2/\lambda R$ may be set equal to its sine and Eq. (6) may be rewritten for the case $\rho = 1$, $\phi = 180°$,

$$|F| \cong \left|2 \sin\left(\frac{2\pi h_1 h_2}{R}\right)\right| = \frac{4\pi h_1 h_2}{\lambda R}.$$

Equation (2) now becomes

$$P_R = P_T G_T G_R \frac{(h_1 h_2)^2}{R^4}. \tag{7}$$

This equation applies to the region lying between the lowest interference lobe and the ray tangent to the earth. At microwave frequencies the lowest lobe lies close to the tangent ray; in consequence, the region in which the equation is valid is small. At lower frequencies the region is greater and accordingly more important. Since it is derived on the assumption of a flat earth, it applies best for short ranges and low antenna heights.[2]

It is of course possible to make all the foregoing considerations more rigorous. Directionality of the transmitter antenna may be taken into account by weighting the amplitudes of direct and reflected rays by the gain factors $f(\theta_1)$ and $f(\theta_2)$, where the $\theta$'s are the respective angles that

---

[1] *Propagation of Short Radio Waves*, Vol. 13, Sec. 5·1, Radiation Laboratory Series.

[2] The validity of Eq. (7) is discussed in detail in *Propagation of Short Radio Waves*, Vol. 13, Secs. 2·13 and 2·15, Radiation Laboratory Series.

the rays make with the axis of the beam. The two rays do not arrive at the receiver from exactly the same direction. The earth is spherical, not flat, and waves reflected from it diverge. All of these considerations give rise to corrections that are small in most cases.[1]

**3·2. The Practical Effects of Interference.**—At microwave frequencies (3000 Mc/sec and up), land reflections are generally considered to be unimportant in producing a regular interference pattern since the surface of the earth is rarely mirrorlike to the degree required for coherent reflection of waves shorter than about 10 cm. Thus microwave early warning radar sets are designed to obtain full coverage without benefit of reflection from the earth. Often only the lowest interference lobe is observed. It has been observed, however, that portable microwave systems located at airports give pronounced interference patterns when their antennas are low (about 4 or 5 ft). This occurs because for such low antenna heights the reflection takes place close to the antenna on the extremely flat surface of the airport. Ground reflection coefficients of more than 0.6 have been observed for 9-cm radiation.

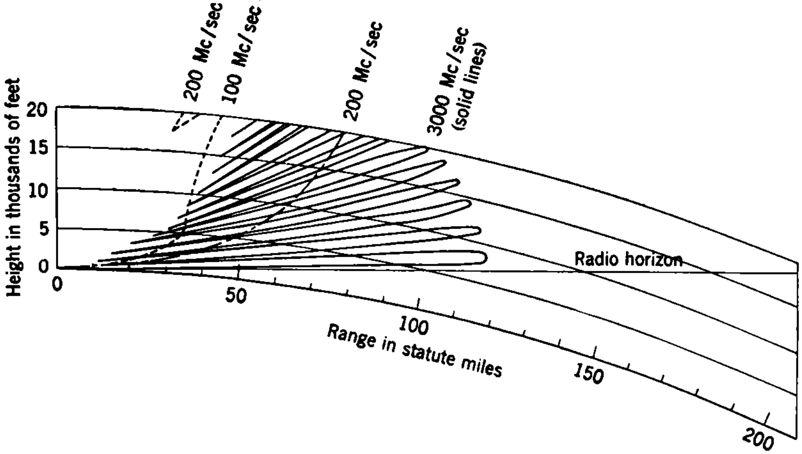


FIG. 3·3.—The lobe structure due to coherent reflection from sea water, as a function of frequency. Contours of equal intensity (−100 db relative to the intensity 1 m from the transmitting antenna) are shown for 100, 200, and 3000 Mc/sec. The transmitting antenna height is 30 ft and polarization is vertical. Refraction is taken into account by assuming a $\frac{4}{3}$ earth radius. Note the great advantage of microwave frequencies in providing coverage at low angles of elevation, near the horizon.

If the antenna heights are greater, and the reflections consequently take place further away and over a larger, rougher area, the reflection may often diminish to the point at which the nulls are negligible.

[1] See *Propagation of Short Radio Waves*, Vol. 13, Secs. 2·13, 5·2, and 5·3, Radiation Laboratory Series.

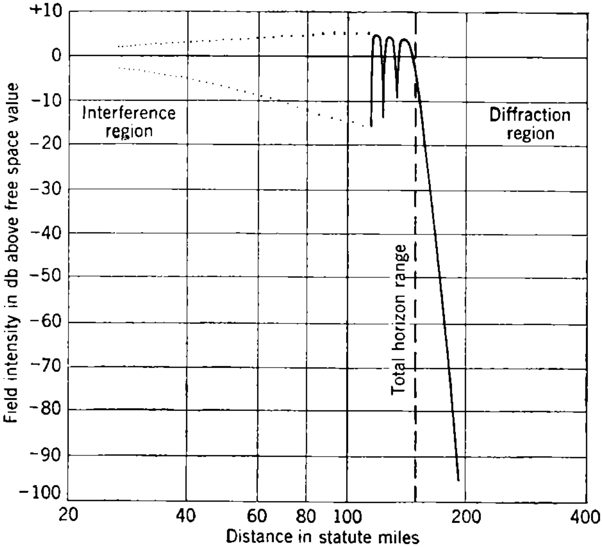


FIG. 3·4.—Field intensity relative to the free-space value. The fluctuations result from reflection from sea water. Data are for a frequency of 3000 Mc/sec, vertical polarization, a transmitter height of 90 ft, and a receiver height of 10,000 ft. Only the furthest lobes are drawn in; the positions of closer maxima and minima are indicated by dots.

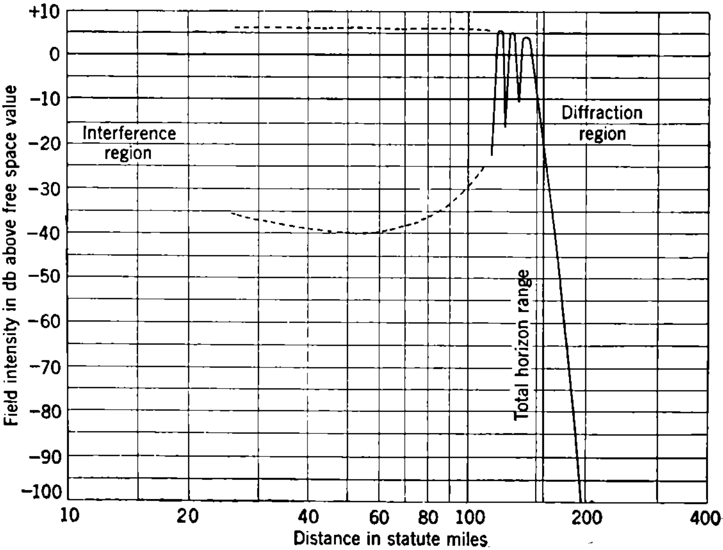


FIG. 3·5.—Field intensity relative to the free-space value as a function of distance from the transmitter. The conditions are the same as for Fig. 3·4, except that the polarization is horizontal. Note the markedly greater depth of the nulls.

*Application of Results.*—One would suppose that, if all quantities were known, they could be substituted into the appropriate equations and would yield correct results. This is true to a degree. The procedure is generally worth carrying out to find the theoretical norm. The results of such calculations are shown for several representative cases in Figs. 3·3, 3·4, 3·5, and 3·6. It is of the greatest importance, however, to realize that such results are to be used with discrimination.

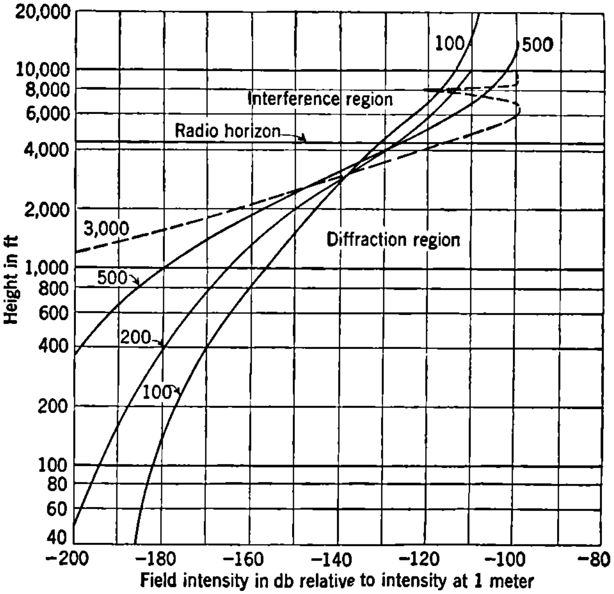


FIG. 3·6.—Field intensity as a function of receiver height in the diffraction region (in the absence of "anomalous" propagation). Transmission is over sea water. The transmitter height is 30 ft, polarization vertical. The values shown are for a distance of 100 miles from the transmitter.

The skepticism implied is not due to any fault of the calculations, but rather to the fact that the conditions of the problem refuse to remain fixed. Neither the atmosphere nor the surface of the earth is constant in its effect. The ground and its covering vegetation change their reflecting properties with their moisture content, and both the ground and the sea change their configurations with the wind. Observations of power received at a fixed point well within the horizon range from a fixed distant transmitter show that very wide fluctuations occur, 30 to 40 db having been observed.

Our picture of the effects of interference, in so far as it concerns system design, should then be about as follows. Within the interference region, under standard refraction conditions, and over a not-too-rough earth, a transmitter at any given height will give rise to a pattern of sharp lobes because of the interference between direct and reflected rays. The height of the maxima will be as much as 6 db above the free-space value, but sharp and deep minima will exist; the intensity may drop almost to zero. The spacing of these maxima and minima will decrease with decreasing wavelength, with distance from the transmitter, and with increasing height above the reflecting surface of either transmitter or receiver. The reflection coefficient of the surface and the refractive index of the atmosphere may vary with time. Thus the depth of the minima and the height of the maxima may fluctuate with time, and their locations may shift.

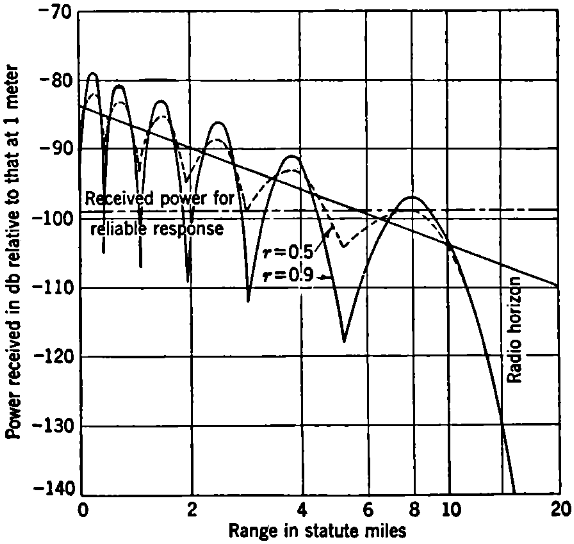


FIG. 3·7.—Graphical evaluation of the effect of interference nulls in the response link of a 9-cm airborne-radar-ground-beacon system. The data are for an aircraft height of 10,000 ft, a beacon height of 5 ft, and horizontal polarization. Note the importance of the value of the reflection coefficient in determining the number and width of the regions in which response failure occurs. The interrogator (AN/APS-2) receiver is exceptionally insensitive because of high duplexing losses.

Furthermore, since the interrogation and reply frequencies of a beacon system are seldom identical, the interference patterns for the two links of a beacon system will be different. If these two frequencies are nearly the same, the respective minima will be close together. The effect will be that of widening the minima somewhat; for an aircraft flying radially one link will fail first, then the other; after the aircraft passes through a minimum, the second link to fail will be the second to recover. With wider frequency separation between the interrogating and reply links the separation of the minima will increase further until they are well resolved, and separate failures of interrogation and reply will be observed. The depth of the minima depends critically upon the reflection coefficient, and thus on the smoothness of the earth and on the polarization and glancing angle of the incident radiation. Whether or not a given minimum will result in the complete loss of the signal is a question of the power margin available and the depth of the null.

A graphical method of evaluating the effect of interference nulls and maxima is shown in Figs. 3·7 and 3·8. These are plots of signal strength vs. range, for a fixed set of system values, for the response link of an airborne-radar–ground-beacon system.

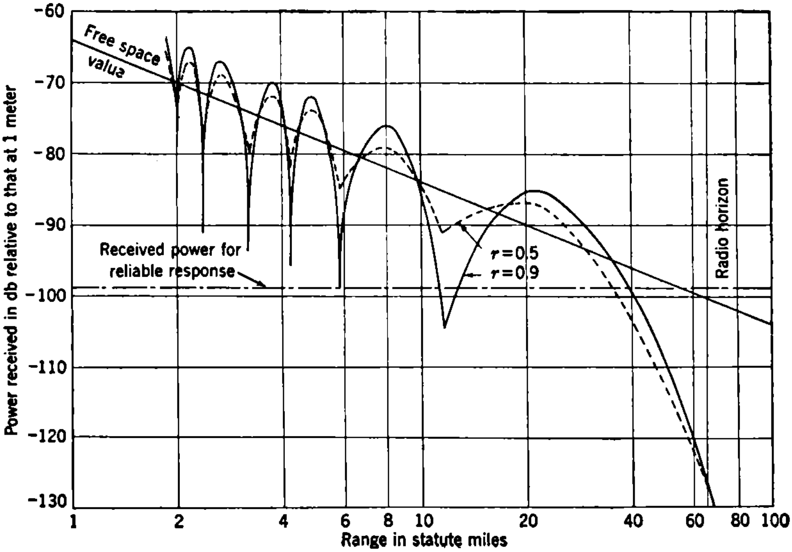


FIG. 3·8.—The same as Fig. 3·7, except that the altitude of the aircraft is 2000 ft.

**3·3. Anomalous Propagation.**—Having discussed the effect of interference in the orthodox case, we must show how atmospheric conditions may drastically alter the picture, especially in transmission over water.

In many cases the atmosphere is in a state in which its index of refraction varies nearly linearly with altitude. It can be shown that propagation in such an atmosphere is equivalent to that in a homogeneous atmosphere about an earth with a radius $ka$, where $k$ is a constant and $a$ the actual radius.[1] The factor $k$ is given by

---

[1] J. C. Schelling, C. R. Burrows, and E. B. Farrell, *Proc. IRE*, **21**, 27 (1933).
C. R. Erglund, A. B. Crawford, and W. W. Mansford, *Bell Syst. Tech. J.*, **14**, 369 (1935).
*Propagation of Short Radio Waves*, Vol. 13, Sec. 2·4, Radiation Laboratory Series.

$$k = \left(1 + a\frac{dn}{dh}\right)^{-1},$$

where $dn/dh$ is the gradient of the refractive index with respect to height above the surface. The use of a fictitious radius is permissible only when the gradient is essentially constant and $k$ is positive. In temperate climates, for altitudes up to several thousand feet, a good average value for $k$ is $\frac{4}{3}$. The value $k = \frac{4}{3}$ is known as the standard value. When the gradient of the index is not nearly linear with height, the concept of an equivalent radius for the earth is no longer valid and the conventional procedures given above for calculating $F$ are no longer applicable.

It is not possible in this book to consider at any length departures of propagation from those of standard refraction. These matters are fully discussed in Vol. 13 of this series, Chaps. 1, 2, and 4. A few recognized cases of this so-called "anomalous" propagation, however, will be pointed out.

1. The gradient of the index of refraction may be lower than standard. This may occur, for example, during the presence of certain types of fog, such as that due to moist, warm air flowing over cold water. In this case, abnormally short ranges may be encountered.
2. A pronounced maximum of the gradient near the ground (a sharp decrease of the index of refraction in the first few hundred feet or less of the atmosphere) results in a layer at the surface with a higher index of refraction than that lying above it, and thus may give rise to the phenomenon generally known as "trapping" or "superrefraction." When such a layer or "duct" exists, radiation starting out nearly parallel to the earth's surface may be partially confined to it; the layer, in effect, acts like a rather leaky waveguide. Extremely long ranges for ground returns may then be obtained by ground and ship radar sets within the duct. This phenomenon may be caused by temperature inversion, sometimes accompanied by a rapid decrease of water-vapor pressure with height. It occurs when offshore winds are warm and dry, or during a sea breeze, or at night during rapid cooling of land masses, and often in winter in polar regions, especially over land.
3. Trapping may also occur when the maximum of the gradient occurs at some higher level. The trapping layer may or may not be ground-based, and may be deeper than in the case described above. The depth affects the maximum wavelength that may be trapped.

The conditions under which anomalous propagation occurs are due to various meteorological factors, and it is the purpose of radar forecasting to study these factors with a view to predicting the occurrence of abnormal conditions. At the moment, it is desirable to emphasize that the usual approximation of $\frac{4}{3}$ $a$ for range calculations is merely a convenient fiction founded on assumptions that are not always valid, and that the existence of many other kinds of atmospheric condition must be taken into account in considering the performance of systems using beacons.

**3·4. Horizon Range.**—We have seen that, for standard refraction, the effect of the atmosphere may be taken into account by assuming that

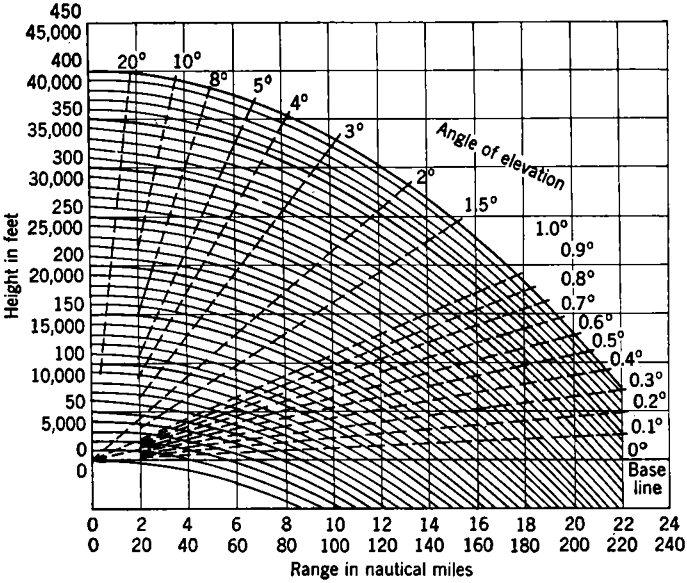


FIG. 3·9.—Horizon ranges and elevation angles as functions of distance and altitude. A $\frac{4}{3}$ earth radius is assumed. Horizon distances for a given altitude are indicated by the intersection of the curve for that altitude with the base line denoting zero elevation. The upper scales for abscissa and ordinates should be read together; likewise the lower scales.

the earth has a radius $\frac{4}{3}$ times its actual value, and by computing geometrical horizon ranges on this assumption. When this is done (as in Fig. 3·9) the horizon will be at a distance $r$ from a point at height $h$ above a spherical earth given by

$$r = 1.22\sqrt{h} \tag{8a}$$

where $h$ is in feet and $r$ is in nautical miles; or by

$$R = \sqrt{2h} \tag{8b}$$

where $R$ is in land miles and $h$ in feet.

The radio horizon distance between a point at height $h_1$ and another at height $h_2$ is

$$r = 1.22\sqrt{h_1} + 1.22\sqrt{h_2}. \tag{9}$$

The square root dependence emphasizes the importance of elevating the antennas of ground beacons as much as possible. Even 100 ft will add 12 nautical miles to the range of an aircraft interrogating a beacon.

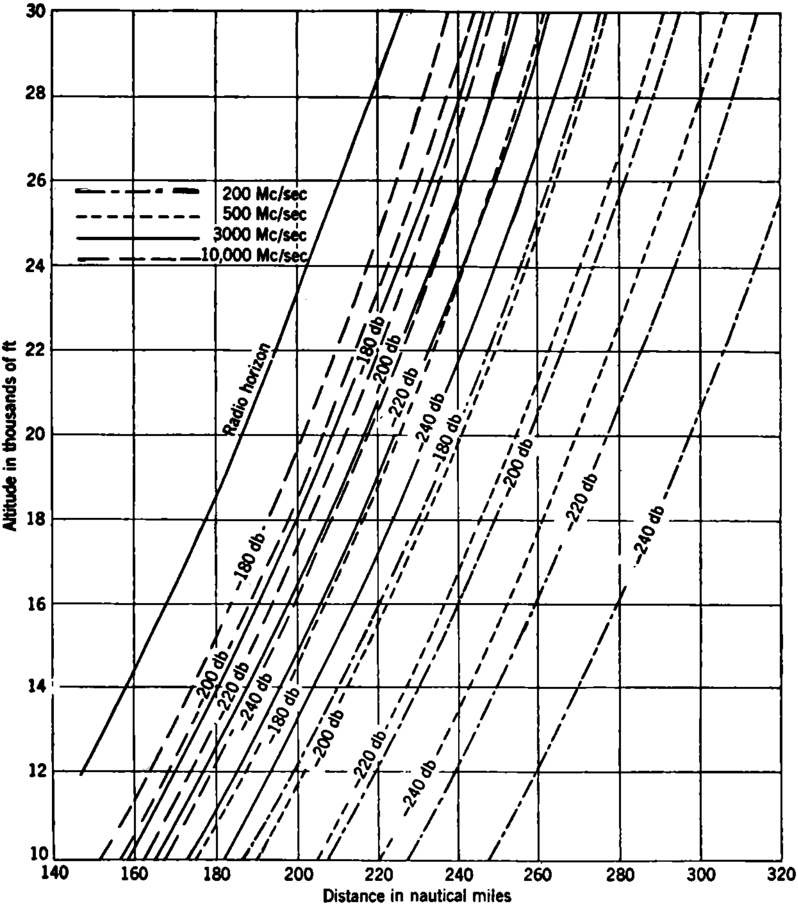


FIG. 3·10.—Frequency and signal strength in the diffraction region. Intensities relative to those at 1 m from the transmitter are plotted for four frequencies as functions of altitude and range. Note the slower decrease in intensity beyond the horizon, at lower frequencies.

**3·5. Diffraction.**—Ranges, of course, are not sharply defined. The intensity of the signal does not drop instantaneously as the horizon is crossed. Near the horizon, diffraction is important. The principal regions of signal coverage are the interference region (within line of sight), a transition region, and the diffraction region beyond the horizon. We have already discussed the interference region. Signals in the diffraction region may be treated theoretically; Fig. 3·10 shows the importance of frequency in determining the signal strength in this region. As might be expected, the shorter the wavelength, the more rapidly the intensity of the signal falls off in the diffraction region.

The effect of diffraction extends into the region within the horizon, as optical theory predicts. In this transition region, theoretical treatment is difficult. A suitable compromise is usually made by joining smoothly the solution for the interference region with that for the diffraction region (see, for example, Fig. 3·7, Sec. 3·2).[1]

Because the intensity of the signal falls off approximately exponentially in the diffraction region, increase in power to increase the range beyond the radar horizon is unprofitable. At 3 cm, for example, 4 db is required for each mile of additional range. This is the reason for calling microwaves "quasi-optical."

On the other hand, atmospheric effects may, by increasing the apparent radius of the earth, or by trapping, occasionally lead to high signal strengths in the diffraction region and to anomalously long ranges. A 10-cm ground beacon has been seen at over 300 miles by an aircraft at 5000 ft. A 200-Mc/sec beacon has been seen at 1500 miles by an aircraft at 5000 ft. In general these effects are much too unreliable for practical use; they should be kept in mind, however, because they occasionally lead to observations on second or third sweeps of the indicator which may be misinterpreted by observers as echoes or beacons in unlikely locations.

## 3·6. Atmospheric Absorption and Scattering.

*Absorption.*—Electromagnetic radiation is subject to absorption by any substance through which it passes. The frequencies at which absorption occurs are characteristic of the substance. In the optical range, they furnish a means of identifying substances.

Frequencies below about 10,000 Mc/sec are not appreciably absorbed in the atmosphere. Two components of the atmosphere, however, do absorb the higher microwave frequencies: oxygen and water vapor.

An absorption curve for oxygen and water vapor is shown in Fig. 3·11. The total absorption by the atmosphere is the combination of these two. The amount of the absorption depends on the position of the path through the atmosphere (which determines the oxygen pressure) and on the vapor pressure and distribution of water vapor along the path. It should be noted that a path tangential to the earth and extending from

---

[1] See also *Propagation of Short Radio Waves*, Vol. 13, Sec. 2·15, Radiation Laboratory Series.

sea level through the entire atmosphere is equivalent to only 125 nautical miles of atmosphere at sea level, as far as oxygen absorption at frequencies off resonance is concerned.

At frequencies above 15,000 Mc/sec, absorption begins to play a more and more important rôle. Absorption differs from other range-limiting factors importantly because it is exponential. The inverse-square law results in a 10-db decrease in intensity between 1 and 10 miles, and again between 10 and 100 miles. An attenuation of 0.5 db per mile, however, will lose only 4.5 db between 1 and 10 miles, but will lose 45 db between 10 and 100 miles. Correspondingly, in this example, a 6-db increase in transmitter power would double the range in the absence of attenuation, for example, from 100 to 200 miles. With 0.5 db attenuation per mile, however, it would add only something under 12 miles of range. The limit thus imposed is almost insurmountable since any large increase in range would require prohibitive increases in power.

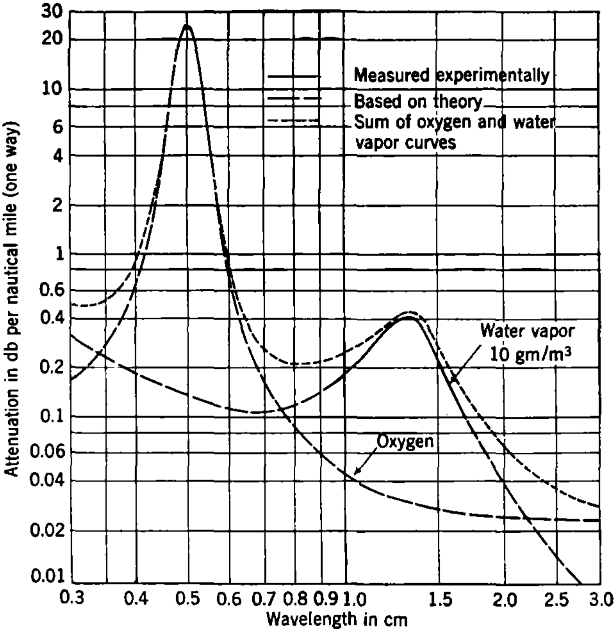


FIG. 3·11.—Atmospheric attenuation due to oxygen and water vapor, at frequencies above 10,000 Mc/sec (wavelengths below 3 cm). The attenuation due to water vapor is usually assumed proportional to its concentration; this is probably true in the range of concentrations encountered in the atmosphere. NOTE: Solid part of each curve has been measured experimentally. Dashed parts of curves are based on theory. Dotted curve is sum of oxygen and water-vapor curves.

Long ranges at frequencies at which appreciable absorption is present are therefore impractically expensive in power; short ranges suffer considerably less. High frequencies may thus be used for short-range systems even when they would be impractical for long-range systems.

At frequencies up to 30,000 to 40,000 Mc/sec, the attenuation due to the oxygen absorption that has its maximum at 50,000 Mc/sec is not very great. However, the water-vapor absorption near 25,000 Mc/sec is important for frequencies above about 18,000 Mc/sec. Because the water-vapor content of the atmosphere is variable, system performance in this region will depend markedly on the atmospheric conditions and become poorer when the vapor pressure is higher. Unfortunately, the need for good performance is greatest when the conditions are least favorable—in bad weather.

Because of atmospheric absorption, frequencies above about 15,000 Mc/sec are not ordinarily useful for horizon-range navigational aids, though for short ranges the limit on frequency can be extended somewhat further.

*Rainstorms.*—In addition to absorption by water in the vapor state, attenuation of signals may occur because of water droplets in the atmosphere.

It is well known that storms and storm clouds are readily observed by microwave radar sets. This property is often useful to pilots for avoiding storm areas. The existence of such storm echoes testifies to the reflection and scattering of microwaves by certain types of cloud. Ordinary clouds consist of extremely minute water droplets that have little effect on signals. The size of droplets that affects signal propagation depends upon the ratio of droplet size to wavelength. In storm clouds the droplets are large enough to give observable reflections.

Such reflection and scattering of radiation necessarily results in attenuation of the signal traversing the cloud. When the attenuation reaches sufficiently high values, radar "shadows" are cast by the cloud, and no echoes from regions behind the cloud are observed.

As we have noted, the effect increases with decreasing wavelength. At 10 cm, cloud shadows are observed only during storms like tropical thundershowers that contain very large drops of water. At 3 cm, somewhat less severe disturbances may produce shadows. At 1 cm, even moderate rainstorms have serious effects.

The actual amount of attenuation caused by such storms is not, ordinarily, very great. It must be remembered that radar shadows from clouds are due to attenuation in two-way transmission. In beacon systems we are concerned with one-way transmission. Experience has shown that beacon operation either at 10 cm or at 3 cm is almost never seriously affected by storms. The existence of attenuation due to storms is, however, one more reason for providing a comfortable margin of power in microwave beacons designed for reliable long-range operation.

**3·7. Polarization.**—The polarization used for a system employing beacons depends on several factors. If the system is one in which the interrogators are used also as radar sets, the considerations that govern the choice of polarization for the radar use will be paramount. As far as the beacon use is concerned, airborne-radar–ground-beacon or airborne-radar–ship-beacon systems will operate satisfactorily with either horizontally or vertically polarized radiation. Over water, the nulls will be less pronounced if vertical polarization is used. For ground and ship beacons, the shorter the wavelength and the higher the beacon antenna installation, the less does polarization matter. Beacons that must operate with both vertically and horizontally polarized interrogating radars may use circularly polarized antennas, with an antenna loss of 3 db in both links of the system.

In microwave ground and ship radar sets, vertical polarization is usually used to minimize interference effects, and horizontal polarization to enhance them; airborne microwave beacons work well enough with both. However, at 10 cm, the smallest antennas for airborne beacons can be made with vertical polarization; they consist of quarter-wave dipoles mounted on the skin of the aircraft.

*Automatic Tracking.*—The conical scan of automatic-tracking sets is sometimes achieved by rotating a suitably designed dipole feed so that the plane of polarization rotates continually. Plane-polarized receiving antennas for beacons will then receive an amount of power proportional to the square of the cosine of the angle between the plane of polarization of the incident radiation and the plane of the beacon antenna. The same will be true on the reply link. Accordingly, the beacon will not be interrogated throughout the entire 360° of rotation of the dipole, but through a smaller angle that depends on the available power. Likewise, the reply signal will be a maximum at points on the cycle 180° apart in phase and zero at some intermediate points. As a consequence, such a system can supply good data either for elevation or for azimuth tracking, but not for both at the same time. Which it will be depends upon the plane in which the beam is offset when the polarizations of the antennas of the radar set and beacon are the same.

To obtain good tracking in both azimuth and elevation, a system must be used in which the amplitude of the received signal depends only upon the orientation of the scanning beam. In a beacon system this is important only for the response link since the beacon response does not vary with the illumination as a radar signal does, but is either present or absent.

This dependence can be achieved in two ways. Either circular polarization can be used for the beacon antennas, or the conical scan can be achieved with a fixed plane of polarization, for example, with an antenna feed nutated around the axis of the paraboloid. Difficulty will be encountered with circularly polarized beacon antennas if the antenna feed of the paraboloid has side lobes of improper polarization since error signals at right angles to the proper direction will then be generated and the tracking will be lost. In practice this has been difficult to correct. Furthermore, the resultant polarization of signals returned by a plane-polarized beacon transmitting antenna mounted on an aircraft is not unique, and is unpredictable. Reply signals are generally elliptically polarized, and the constants of the ellipse are different for every aspect of the aircraft. Accordingly, attempts to radiate circularly polarized signals from an airborne beacon are not likely to be successful.

## COVERAGE CONSIDERATIONS

The coverage of a radar system or of a beacon system refers to the shape and volume of the space in which useful signals are obtained. The extent of the coverage depends upon the range of the system and upon the antenna patterns. The required coverage is ordinarily one of the fundamental parameters determining the design of the system.

In systems used both for radar echoes and for beacon replies, the requirement for beacon replies is usually that the beacon coverage be at least coextensive with the radar coverage, and greater if possible. Unless the beacon coverage is at least the same as the radar coverage, there will be regions in which radar signals will not be accompanied by beacon signals. Since one of the chief advantages of using beacons is their extension of the range, it is most often desirable to have beacon coverage greater than radar coverage.

The shape of the coverage volume is largely determined by the antenna patterns of interrogator and beacon antennas. We will consider various types of antenna and the problems arising in connection with them.

**3·8. General Considerations.**—The coverage of a radar set is determined by the antenna pattern. It determines the illumination of the target and the amount of the reflected signal received. Thus, the two-way or signal pattern is the square of the one-way or illumination pattern. The signal pattern is, therefore, sharper and narrower than that for the one-way illumination. For the beacon coverage to equal or exceed the radar coverage, the receiver sensitivity and receiving antenna pattern of the beacon must be adequate to trigger a response that will return at least as much energy to the interrogator receiver as would a target of appropriate size at the same place. This requirement is to be fulfilled at maximum radar range. It cannot possibly be fulfilled at all ranges because the radar return varies as the inverse fourth power of the distance, and the beacon return varies as the inverse square. There will, accordingly, always be a certain range beyond which the beacon return exceeds the radar echo and within which the reverse is true.

The importance of this consideration depends upon the display method used. It is of little consequence if beacon returns and radar returns are not displayed simultaneously since in this case the radar and beacon signals are, by definition, under independent control. When the returns are displayed simultaneously, however, independent control is of great importance for getting satisfactory display. The use of independent receivers and independent video-limit levels is clearly indicated.[1]

Further requirements must be imposed upon the beacon antenna patterns. The patterns of the receiving and transmitting antennas of the beacon must be such that when the beacon is within the coverage volume it will at all times receive enough energy for triggering and return enough for adequate display. In general, then, the patterns of the beacon antennas must be less directional than those of radar antennas.

**3·9. Airborne Interrogator—Surface Beacon.**—The coverage volume of an airborne radar set depends upon the altitude of the aircraft. Both maximum and minimum ranges of objects on the ground will vary with altitude. Depending upon the type of antenna used, there will be a cone of greater or lesser vertex angle under the aircraft, in which the signal intensity will be very low. (See Fig. 3·12.) When the radar antenna is designed for uniform ground return (cosecant-squared antenna) this cone is small. When the radar fails to trigger a ground beacon within this cone, a so-called "cone of silence" exists for the radar set. A ground beacon also has a cone of silence directly above it, unless special care is taken to illuminate this region. The combination of patterns of radar and beacon often results in no signal being received by an airborne radar close to a ground beacon. The difficulty of eliminating the region of silence, which naturally increases in diameter with the altitude, becomes so severe at high altitudes that most of the energy of the beacon would have to be directed upward to insure complete coverage. Accord-

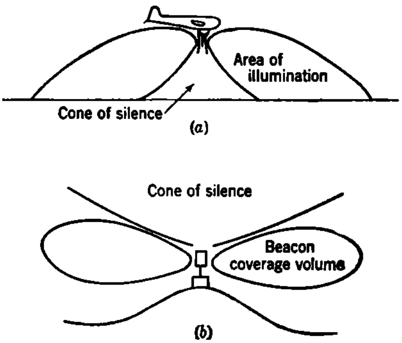


FIG. 3·12.—The "cone of silence" (a) for an airborne radar set; (b) for a ground beacon.

---

[1] See, in this connection, Chap. 17, Sec. 17·3.

ingly, the attempt is usually given up and a compromise by which the cone of silence is kept reasonably small is worked out. In practice, the cone of silence above a beacon is of little navigational significance since there is no directionality in azimuth vertically above the beacon. Only at low altitudes is it likely to be important to know that one is directly over a beacon and we have seen that response can usually be maintained at low altitudes.

*Shipborne Beacons.*—To maintain good operation of shipboard beacons, the effect of the shifting of the antenna pattern that results from the motion of the ship must be counteracted. This can be done in either of two ways: At the expense of antenna gain, the elevation width of the pattern can be made so broad that roll and pitch do not exceed the vertical beamwidth, or the beacon antennas may be stabilized vertically by a gyrostabilizer.

**3·10. Ground Radar—Airborne Beacons.**—Good angular coverage of most of the sky can be obtained with ground radar. The airborne beacon antenna must accordingly be omnidirectional, with a fairly broad vertical beam to take into account the banking of the aircraft.

At microwave frequencies, difficulty arises because it is impractical, if not impossible, to install omnidirectional airborne antennas that will not be shielded from the ground radar during maneuvers of the aircraft. Good coverage is thus limited to the intervals in which aircraft are not executing maneuvers that result in shielding the antenna on the aircraft from the radar on the ground. In practice this problem is not serious because the identity of the beacon return is unmistakable and the beacon is picked up again as soon as the maneuver is over. The worst case of this type occurs with automatic-tracking radar sets, in which tracking will be thrown off and must be locked on again after the signal returns.

In antenna installations for airborne beacons, reflections and diffraction from the surfaces of the aircraft are important in determining the over-all antenna pattern. Accordingly, every actual installation must be tested carefully since theoretical patterns are too hopelessly complicated to permit evaluation.

**3·11. Example of Range Calculations.**—A sample calculation of range will show how all the factors influencing range may be taken into account.

Suppose an airborne radar of light weight has been designed, together with a suitable ground beacon, as a general-purpose navigational aid. Given the characteristics of the two sets, range may be calculated.

The pertinent characteristics for this calculation are set forth in Table 3·1.

To evaluate the effect of coherent interference, it will be convenient first to calculate the power available at the input terminals of the beacon receiver at a range of 1 nautical mile. For this purpose, the range equa-

TABLE 3·1.—CHARACTERISTICS OF A 3-CM AIRBORNE RADAR AND A GROUND BEACON

| | Radar | Beacon |
|---|---|---|
| Frequency: interrogation | 9330–9420 Mc/sec | |
| Frequency: response | | 9310 Mc/sec |
| Pulse power (kw) | $P_T = 8.0$ | $P'_T = 0.3$ |
| Antenna gain (max) | $G_T = G_R = 700$ | $G'_T = G'_R = 10$ |
| Beamwidth (half power) | 5° (azimuth) | 12° (elevation) |
| Scan | 360° | |
| Antenna rotation speed, rpm | 20 | |
| Pulse length, μsec | 2.5 (beacon) | 0.5 (3, with video-stretching) |
| Sweep (max) | 100 miles | |
| Display | 5-in. PPI | |
| Receiver sensitivity, watts | $5 \cdot 10^{-13}$ (min. detectable signal) | $P'_R = 1 \cdot 10^{-9}$ (triggering) |
| R-f line loss (receiving) | $L_R = 5$ db max. | $L'_R = 0.5$ db |
| R-f line loss (transmitting), db | $L_T = 0.5$ | $L'_T = 0.5$ |

tion in the form of Eq. (4), Sec. 2·5 is used. This equation reads

$$10 \log_{10} P'_R = 10 \log_{10} \frac{P_T G_T G'_R \lambda_i^2}{(4\pi R_i)^2} - (L_T + M_T + A_i + M'_R + L'_R).$$

It gives the power available in decibels below 1 watt.

We assume that $M_T$, the pattern loss of the interrogator transmitting antenna, is 3 db because we demand triggering between half-power points; we also assume that $M'_R$, the pattern loss of the beacon receiving antenna, is zero at long range and that atmospheric absorption $A_i$ may be neglected. All other relevant quantities are given in Table 3·1. Substituting them in the above equation, with $R_i$ taken as 1 nautical mile, we find that

$$10 \log_{10} P'_R = -44.$$

The power available at the input terminals of the beacon receiver at a range of 1 mile is then 44 dbw.

A curve showing the received power for free-space conditions can now be drawn, as shown in Fig. 3·13; it shows the inverse square drop. It intersects the beacon-triggering level, which is at 89.5 dbw, at a range of 185 miles. This is the *free-space range* for interrogation.

A similar calculation is now made for the response link. For this link, the relevant equation is Eq. (5), Sec. 2·5, which reads

$$10 \log_{10} P_R = 10 \log_{10} \frac{P'_T G'_T G_R \lambda_r^2}{(4\pi R_r)^2} - (L'_T + M'_T + A_r + M_R + L_R).$$

We assume that $M'_T$ is zero, $A_r$ is zero, and $M_R$ is 3 db. For $R_r$ equal to 1 nautical mile, this gives

$$10 \log_{10} P_R = -62.5.$$

The power at the input terminals of the interrogator receiver is thus 62.5 dbw. We have not yet taken into account display losses. The receiver sensitivity quoted, $5 \times 10^{-13}$ watt, is for a minimum detectable signal on the PPI.

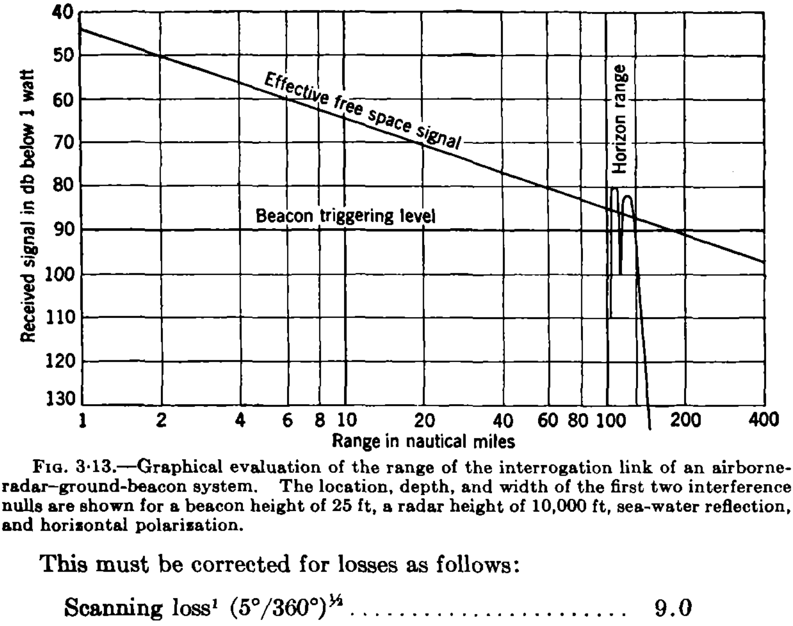


FIG. 3·13.—Graphical evaluation of the range of the interrogation link of an airborne-radar–ground-beacon system. The location, depth, and width of the first two interference nulls are shown for a beacon height of 25 ft, a radar height of 10,000 ft, sea-water reflection, and horizontal polarization.

This must be corrected for losses as follows:

| | |
|---|---|
| Scanning loss[1] $(5°/360°)^{½}$ | 9.0 |
| Long-sweep loss,[2] from Fig. 2·1, Sec. 2·6 | 2.0 |
| Total | 11.0 db |

The receiver sensitivity useful for reliable beacon signals is then 112 db below 1 watt, $-5 \times 10^{-13}$ wat tor 123 db below 1 watt, corrected for 11-db losses. This yields a calculated free-space range for response of 300 miles (see Fig. 3·14).

The system is not quite balanced. The response link has 4 db more power margin than the interrogation link; it is 4 db stronger. In general,

---

[1] The scanning period is so short (3 sec) that this expression may be used without correction.

[2] The pulse length on the sweep is 3 μsec · 60 mm/100 mile · 12 μsec per mile = 0.15 mm. Note that without video-stretching the ½-μsec beacon pulse would be 0.025 mm long, and the loss would then be 5 db (compared with optimum conditions).

a slight unbalance in this direction is a good thing: when the beacon signal appears, it appears strongly, and less fatigue and strain are necessary to pick it up at long range. The range of the system will depend less on the skill of the radar operator than it would if the system were unbalanced in the opposite sense.

The effect of nulls resulting from reflections over water can now be evaluated. In Fig. 3·13 the first two interference maxima and minima have been sketched in for a beacon 25 ft high, a radar set at 10,000 ft, and horizontal polarization. The nulls are narrow but deep. At the beacon-triggering level the farthest (and widest) null is less than 2 miles wide. For general navigational purposes such nulls are of little importance since they are narrow. Neither pickup of the beacon at long range nor ordinary navigation with respect to it are seriously affected by failure to see the beacon for a half-dozen scans, more or less.

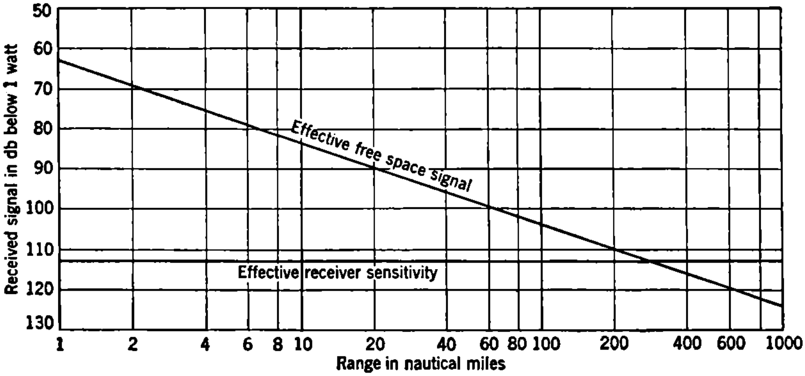


FIG. 3·14.—Graphical evaluation of the range of the response link. The effects of interference are not shown for this link; they are evaluated in the same way as that shown for Fig. 3·13.

The great increase in power level necessary to override the nulls completely at long range can also be evaluated from such diagrams. More than 20 db would be required in our case; for interrogation this might mean 10 db more in transmitted power and 10 db more sensitivity in the beacon receiver. The advantage of vertical polarization is likewise evident (see Figs. 3·4 and 3·5).

The peculiar advantages of microwaves are also evident. The fraction of time spent in nulls is about the same for all frequencies, for systems of equal range. At microwave frequencies the nulls are narrow and closely spaced; at lower frequencies they would be wider and farther apart. At 30 cm, the null under our conditions would be 10 to 20 miles wide, and thus of considerably greater navigational significance, especially over water.

## CHAPTER 4

# FREQUENCY CONSIDERATIONS

By A. Roberts

The choice of the frequencies for a beacon system is dictated by a variety of considerations. The primary question is whether the system is to be used for radar echoes as well as for beacon responses. If it is, the considerations that determine the radar frequency are likely to be paramount. If the frequency thus determined is compatible with the conditions governing beacon-frequency choice, the radar set can be used to trigger the beacons and to receive their responses. If not, separate interrogator-responsors with a common display system should be considered.

## GENERAL CONSIDERATIONS IN THE CHOICE OF FREQUENCY

**4·1. Omnidirectional Systems.** *Free-space Systems.*—Let us calculate the power $P'_R$ received by an omnidirectional beacon antenna of gain $G'$, from an interrogator of power $P_T$, with an omnidirectional antenna of gain $G$. Using the notation of Sec. 2·2, and assuming free-space conditions,

$$P'_R = \frac{P_T G G' \lambda^2}{(4\pi R)^2}. \tag{1}$$

Solving for $R$,

$$R = \frac{\lambda}{4\pi} \left( \frac{P_T G G'}{P'_R} \right)^{\frac{1}{2}}. \tag{2}$$

For antennas of fixed gain at both ends of the system (for example, antennas omnidirectional in azimuth with a fixed vertical beamwidth), the range is proportional to the wavelength. Accordingly, systems in which *range only* is desired should use long wavelengths, not microwaves.

*Over-water Systems.*—An exception to this conclusion occurs in the use of ship-to-ship or ship-to-shore systems, which may operate all or most of the time in the region below the first interference lobe described by Eq. (7), Sec. 3·1,

$$P'_R = P_T G_T G'_R \frac{(h_1 h_2)^2}{R^4}. \tag{3}$$

Solving for $R$,

$$R = (h_1 h_2)^{\frac{1}{2}} \left( \frac{P_T G_T G'_R}{P'_R} \right)^{\frac{1}{4}}. \tag{4}$$

For such systems, the range is independent of wavelength when omnidirectional antennas of fixed gain (for example, dipoles) are used. The choice of frequency is then to a first approximation immaterial as long as the frequency is not chosen so high that the system no longer operates in the region described by Eq. (3).

The implications of Eq. (4) for ship-to-shore beacon systems are quite important. In the first place, the range follows an inverse fourth-power law rather than the inverse-square law. Second, it places great emphasis on the heights $h_1$ and $h_2$ of the two ends of the system; doubling either height results in a fourfold increase in received power. Third, the received power is independent of wavelength for transmission between antennas of fixed gain. This is in marked distinction to the free-space equation [Eq. (2)] where the power received is proportional to wavelength for transmission between antennas of fixed gain. In ship-to-shore beacon systems, for example, transmission between dipoles gives results independent of the wavelength. Omnidirectional ship-to-ship and ship-to-shore beacon systems may profitably employ considerably higher frequencies than systems in which the range is governed by Eq. (2).

*Examples.*—An example of an omnidirectional range-only system is one in which a nondirectional airborne interrogator is used with nondirectional ground beacons as a range-only auxiliary to radio ranges and landing systems in which fixed aircraft courses determined by other means are being followed. Another example is special equipment for determining position accurately by means of precision range measurements on two or more beacons at known locations (triangulation systems). Still another would be nondirectional survey beacons for measuring the distance between points on the ground. In such systems, the provision of means for identifying the beacons is important to prevent confusion.

*Applications.*—Range-only systems are obviously inferior for identification purposes and particularly for identification of aircraft. Correlation with radar reflections is difficult or impossible except when traffic density is very small.

Omnidirectional systems should be low-power or short-range or both; otherwise, useless interrogating signals will fill large volumes of space. Omnidirectional systems that do not, in fact, have to receive signals from widely different directions are uneconomical. The emission of signals into regions where they are not needed is wasteful and produces unwarranted interference.

**4·2. Directional Interrogation, Omnidirectional Response.** *Free Space.* Consider now an omnidirectional beacon being interrogated by an interrogator with a directional antenna of fixed aperture $A$. What is the dependence of range on wavelength?[1]

---

[1] Identical considerations will apply to a system with omnidirectional interrogation and a directional reply.

We have from Eq. (1), neglecting losses, for free-space conditions,

$$P'_R = \frac{P_T G' G \lambda^2}{(4\pi R)^2}.$$

But now

$$G = \frac{4\pi k A}{\lambda^2}, \tag{5}$$

with $k \approx 0.6$ to $0.7$, and $A$ the area of the transmitting radiator. Thus

$$R = \left( \frac{P_T G' \cdot kA}{4\pi P'_R} \right)^{\frac{1}{2}}, \tag{6}$$

which is *independent* of the wavelength but proportional to the linear dimensions of the transmitting radiator. This has been shown for the interrogating link only; it is obviously true as well for the response link.

Accordingly, when a directional interrogator and an omnidirectional beacon are to be employed, the range of the system is *to a first approximation independent of wavelength.*

Since the aperture $A$ is fixed, $G\lambda^2$ will remain constant, $G$ increasing as $\lambda^2$ decreases. The beamwidth of the interrogator antenna thus decreases linearly with the wavelength, and shorter wavelengths will result in sharper beams and better azimuth resolution.

Since the range of the system is independent of wavelength, and since antenna aperture is one of the limiting factors in airborne-radar design, the beacon places no limitation on the choice of frequency for airborne radar, to a first approximation. Other considerations, however, do enter. These are the maintenance of the required receiver sensitivity and power output for the frequency required; the difficulties of using very sharp beams in aircraft, which roll, pitch, and yaw; the scanning losses introduced by narrow beams; and the maintenance of a sufficiently narrow frequency "scatter band." The attenuation by the atmosphere of frequencies above 15,000 Mc/sec must also be considered.

This approximate independence of wavelength is one of the considerations that make the use of radar sets as interrogators for beacons attractive. It results in maximum economy of equipment, in the automatic provision of beacon information as accurate as the radar information, and in extremely simple and direct correlation of beacon and radar information.

*Over Water.*—Ship-to-ship and ship-to-shore systems operating in the region below the first interference lobe, again, are exceptions to the conclusion that the range of the system is independent of wavelength. Equation (4) now shows that for a system with an antenna of fixed aperture at one end and an antenna of fixed gain at the other, the range is proportional to the square root of the frequency. For such systems, the highest practical frequency should be chosen; this gives not only the maximum range but also the best azimuth resolution. The highest practical frequency will be determined by the values of $h_1$ and $h_2$ likely to be encountered, by the range requirements, by considerations of atmospheric absorption, and by practical considerations such as availability of suitable equipment.

**4·3. Interrogators as Radar Sets.**—One problem still remains. Suppose it is desired to design a beacon system to give both range and azimuth information, to operate independently of any radar system. How shall its frequency be determined?

Before answering this question, let us first note the following. A system with good azimuth resolution will be able to function as a radar set with results similar in precision to those obtained by the use of beacons. It is then pertinent to ask what the radar range of such an interrogator will be and how it will compare with the beacon range.

Before we attempt a quantitative solution to this question, we can first make some qualitative observations. Clearly, if the beacon is less sensitive and emits less power, the interrogator must be more powerful and have a more sensitive receiver; the performance of the interrogator as a radar set will be better. Thus, if low-performance beacons (that is, those with low sensitivity and low power output) must be used for reasons of economy of weight or for any other reasons, the interrogator must have better potential radar performance than it would if the beacon could be more sensitive and emit more power. Likewise, the greater the required beacon range and the greater the safety margin of excess signal required, the more powerful the interrogator must be. It is evident that systems that are required to have a large margin of safety at horizon range at all altitudes are likely to have good radar performance. This fact is relevant particularly for the design of systems for precision navigation. The quantitative treatment will show that for representative values of the system parameters, the radar range of microwave interrogators is about one-tenth to two-tenths of the beacon range.

The quantitative treatment is based on the range equation. We assume for convenience a balanced system with equal interrogation and response ranges. For radar, the received (reflected) signal at the range $R_r$, from a target of cross section $\sigma$, will be $P_r$, given[1] by

$$P_r = \frac{P_T \sigma k^2 A^2}{4\pi\lambda^2 R_r^4}. \tag{7}$$

For beacon operation at the range $R_b$, the power received from the beacon is (in the notation of Chap. 2)

---

[1] *Radar System Engineering*, Vol. 1, Sec. 2·4, Eq. (4*a*), Radiation Laboratory Series.

$$P_b = \frac{P'_T G'_T k A}{4\pi R_b^2}. \tag{8}$$

At the same range $R_r = R_b = R$; then, from Eqs. (5), (7), and (8)

$$\frac{P_b}{P_r} = \frac{P'_T G'_T \lambda^2 R^2}{P_T k A \sigma} = \frac{P'_T G'_T 4\pi R^2}{P_T G \sigma}. \tag{9}$$

This equation gives the ratio of beacon signal to radar signal at any range $R$. This ratio will be unity (that is, equal signals) at the critical range $R_{\text{crit}}$ when

$$P_T G \sigma = P'_T G'_T \cdot 4\pi R_{\text{crit}}^2$$

or

$$R_{\text{crit}} = \left(\frac{P_T G \sigma}{4\pi P'_T G'_T}\right)^{\frac{1}{2}}. \tag{10}$$

The maximum radar range of the beacon interrogator is given by that range $R_r$ at which the return echo is equal to the minimum useful power $P_R$. This is given by the relation[1]

$$P_R = \frac{P_T \sigma G^2 \lambda^2}{(4\pi)^3 R_r^4}, \tag{11}$$

whence

$$R_r = \left[\frac{P_T \sigma G^2 \lambda^2}{(4\pi)^3 P_R}\right]^{\frac{1}{4}}. \tag{12}$$

For the range of the system for beacon responses, we have, for the interrogating link, from Eq. (2)

$$R_b = \left[\frac{P_T G G'_R \lambda^2}{(4\pi)^2 P'_R}\right]^{\frac{1}{2}}. \tag{13}$$

The ratio of radar to beacon range is then

$$\begin{aligned}\frac{R_r}{R_b} &= \left\{\frac{P_T \sigma G^2 \lambda^2 [(4\pi)^2 P'_R]^2}{(4\pi)\ P_R (P_T G G'_R \lambda^2)^2}\right\}^{\frac{1}{4}} \\ &= \left(\frac{4\pi \sigma P'^2_R}{P_R \lambda^2 \cdot P_T G'^2_R}\right)^{\frac{1}{4}}.\end{aligned} \tag{14}$$

For a balanced system the power-product theorem [Eq. (6) Sec. 2·10] gives us $P_R P_T = P'_R P'_T$; then, if $G'_R = G'_T = G'$,

$$\frac{R_r}{R_b} = \left(\frac{4\pi\sigma}{G'^2 \lambda^2} \frac{P'_R}{P'_T}\right)^{\frac{1}{4}}. \tag{15}$$

The ratio increases with decreasing wavelengths as $\lambda^{-\frac{1}{2}}$. Furthermore, for a balanced system, the ratio depends only on the properties of

---

[1] *Radar System Engineering*, Vol. 1, Sec. 2·3, Eq. (3*b*), Radiation Laboratory Series.

the beacon except for the obvious dependence on $\sigma$. It is greatest for small beacons (low power, low sensitivity). Unless the system is balanced, Eq. (14) should be used rather than Eq. (15); otherwise the value of $P'_T$ to be used in Eq. (15) should be the one that would make the system balance.

*Examples.*—For a typical airborne-beacon-ground-radar system, with the values $\sigma = 10^5$ cm², $P'_R = 5 \times 10^{-8}$ watt, $P_T = 5 \times 10^4$ watts, $P_R = 5 \times 10^{-12}$ watt, $\lambda = 10$ cm, $G' = 1$, $G = 10^3$, Eq. (14) gives

$$\frac{R_r}{R_b} = 0.12.$$

The range for radar echoes will be, from Eq. (12),

$$R_r = 23 \text{ km}.$$

This predicts a beacon range of 23/0.12 = 200 km. Let us verify this. From Eq. (13)

$$R_b = \frac{\lambda}{4\pi}\left(\frac{P_T G G'}{P'_R}\right)^{1/2} = 200 \text{ km}.$$

In airborne-interrogator-ground-beacon systems, shorter wavelengths and larger values of $\sigma$ for ground reflections will tend to compensate for the higher values of $G'$ and for smaller values of $P'_R/P'_T$. A typical case might, for example, have $\sigma = 5 \times 10^7$ cm², $P'_R = 5 \times 10^{-9}$ watt, $P_T = 1 \times 10^4$ watts, $P_R = 5 \times 10^{-12}$ watt, $\lambda = 3.2$ cm, $G' = 10$, $G = 10^3$. Equation (14) now gives

$$\frac{R_r}{R_b} = (3 \times 10^{-4})^{1/4} = 0.13.$$

The radar range will be, from Eq. (12)

$$R_r = (5 \cdot 10^{26})^{1/4} = 47 \text{ km}.$$

The beacon range should be 47/0.13 = 360 km. We get from Eq. (13)

$$R_b = \frac{1}{4}\left(\frac{10^{17}}{5}\right)^{1/2} = 350 \text{ km}.$$

These examples demonstrate the principle that highly directional interrogator systems are in general useful as radar systems and should be designed as such.

*Choice of Frequency.*—We may now answer the question raised above as to wavelength dependence in systems that are to give good azimuth information but are intended for beacon operation only. If it should be decided not to use the system for reflections, the governing equation is

Eq. (6); then the wavelength should be the longest that will give the required resolution with an antenna of maximum permissible area $A$. We have seen that such a design is uneconomical for airborne use and of doubtful value for ground use.

We see that, as would be expected, interrogator-responsors are progressively more useful as radar sets, as the beacon-transmitter power and receiver-sensitivity decrease, and as $\lambda$ decreases; for horizon- or greater-range beacon systems they are at least respectable as radar systems. Thus for good angular resolution, which implies high gain, radar interrogators ought to be used. One implication of this is that systems using crystal-video beacon receivers should, in general, have radar interrogators; the converse is that systems for beacon use alone should have superheterodyne beacon receivers. The most appropriate system of this nature, and the most flexible too, will be one in which the beacon transmitter power and receiver sensitivity are identical with the power and sensitivity of the interrogator. In such a system the same equipment can often be used both as a beacon and as an interrogator by means of suitable switching circuits.

We may conclude also that a system for accurate range and bearing should usually have a radar interrogator. This applies to all ground and ship installations working with airborne beacons, all the more because the latter should, in general, be lightweight and thus low-power. Airborne low-power interrogators, for range only, should not be microwave; airborne interrogators with narrow beams should indeed be radar sets.

Even in the special case of long-range airborne interrogators for range only, it appears wise to consider the alternative use of radar interrogation. Here the improved radar performance, the minimizing of interference by using narrow beams, and the navigational facilities afforded by radar may make the choice preferable. In view of the amount of computing equipment needed for triangulation and the development of lightweight, low-drag, microwave scanning antennas and mounts, the microwave alternative in this case may be preferable to an uhf system.

## SYSTEM CONSIDERATIONS IN THE CHOICE OF FREQUENCY

**4·4. Ground or Ship Beacons.**—Ground or ship beacons can be designed for any useful portion of the radio spectrum. Their size, complexity, and cost are a function of many variables, including frequency.

In a ground beacon for aircraft navigation, the maximum range is horizon-limited if enough power and sensitivity are available. This is technically feasible for reasonably wide frequency bands at all useful frequencies. The horizon range at a given altitude varies somewhat with frequency above 100 Mc/sec, but only slightly (see Fig. 3·9).

Since ground-beacon antennas will generally be nondirectional in azimuth, the useful gain in the vertical plane will be limited by the minimum beamwidth and the beam shape necessary for good coverage at all altitudes. In shipboard antennas, the gain will be limited by roll and pitch—unless the antenna is stabilized, in which case the situation is identical with that for ground beacons. The antenna cross section will decrease with increasing frequency after the minimum beamwidth has been reached. Maintenance of the same range then requires more receiver sensitivity in the beacon or more received power per unit area from the interrogator.

**4·5. Airborne Beacons.**—We can apply the same considerations to airborne beacons. Here, however, considerations of aircraft motion lead us to the conclusion that in order to permit banking to occur without loss of the signal, the beamwidth of the beacon antenna must not be less than some fixed angle that is about 45°. As with ground antennas, a maximum *constant* antenna gain is required. Again the range is to a first approximation independent of wavelength for a fixed beacon-antenna gain and a fixed size of the antenna of the ground radar.

With airborne beacons, however, another consideration enters. The problem of mounting an omnidirectional antenna on an aircraft in such a way as to provide good coverage at all azimuths becomes progressively more difficult as the frequency increases. At 200 Mc/sec, omnidirectional quarter-wave dipoles will provide such coverage with relative ease. At 1000 Mc/sec, the problem is more difficult, but soluble. At 3000 Mc/sec, the task is still more difficult. Solutions can be found, in many cases, only by designing special antennas for a particular aircraft. The problem has not been attempted at 3 cm as yet.

The reason for the difficulty is, of course, the increasing ratio of aircraft dimensions to wavelength as the frequency increases. The larger the ratio becomes, the greater becomes the likelihood of occurrence of deep nulls resulting from interference between waves that have traversed paths differing by large fractions of a wavelength. This consideration militates against the choice of wavelengths of less than 10 cm for airborne beacons.

**4·6. Crossband-beacon Systems.**—In the type of beacon system usually referred to as "crossband," the interrogation and reply links are on widely different frequencies; in general, separate transmitting and receiving antennas are used at each end of the system. For this case the range equation takes the form

$$\begin{aligned} R_i &= \frac{\lambda_i}{2\pi}\left(\frac{P_T G_T G'_R}{P'_R}\right)^{1/2}, \\ R_r &= \frac{\lambda_r}{2\pi}\left(\frac{P'_T G'_T G_R}{P_R}\right)^{1/2}, \end{aligned} \tag{16}$$

where the subscripts for $i$ and $r$ refer to the interrogation and response links respectively. The power product theorem in the form of Eq. (6), Sec. 2·10, is clearly inapplicable even for a balanced system. We may rewrite it in the form

$$\lambda_i^2 \frac{P_T P_R G_T}{G_R} = \lambda_r^2 \frac{P'_T P'_R G'_T}{G'_R}. \tag{17}$$

Since the beacon antennas are omnidirectional, cannot have high gain, and should have identical coverage, considerations of economy of beacon size would clearly dictate that $\lambda_r > \lambda_i$. Economy of interrogator-transmitter power dictates a large value of $G_T/G_R$; it also favors $\lambda_r > \lambda_i$.

A crossband system with $\lambda_r > \lambda_i$ thus yields the maximum economy of equipment in the interrogation and beacon sets. Let us now consider the operational characteristics of such a system.

*Crossband Systems with* $\lambda_r > \lambda_i$.—A crossband system with a higher frequency for the interrogation link than for the response link has these characteristics:

1. The angular width of the beacon arc on the display screen is controlled by the interrogating link. It is the angle over which the signal from the interrogator will trigger the beacon during a scan. This follows from the fact that $G_T$ will undoubtedly exceed $G_R$ (for equal effective areas $G_T/G_R \sim \lambda_r^2/\lambda_i^2$).
2. The angular width of the reply will vary with range in the usual way—that is, as determined by the pattern of the interrogator transmitting antenna—and side-lobe triggering will occur at close ranges.
3. In contrast to the control possible with high-gain receiving antennas, reducing the gain of the receiver in the interrogator will not eliminate the side lobes unless $G_R \approx G_T$.
4. Accordingly, arc-width control cannot be achieved except by auxiliary means. These include reducing interrogator-transmitter power and varying the interrogation-code parameters to reduce the effective beacon-receiver sensitivity.
5. The angular resolution over the largest fraction of the range will, however, be much better than it would be if $\lambda_r$ were used for both links.
6. Somewhat better continuity of response from an airborne beacon may be achieved, perhaps, in a crossband system than in a microwave system. As we have seen in Sec. 4·5, it is more difficult to achieve a satisfactory coverage pattern for the transmitting antenna of an airborne beacon if the transmission is at microwave frequencies than if it is at lower frequencies. This is of importance

only in the reply link; for interrogation the systems are identical and can both be improved to any desired degree of excellence by studding the aircraft with as many receiving antennas as may be desired. To avoid interference effects each must have its own crystal-video receiver; the outputs of these must all be connected together.

We see that a simple crossband system of this type offers economy of construction as compared with straight microwave construction. For azimuth information it is superior to nondirectional systems, and probably superior to lobe-switched directional systems; but it is not equal in performance to complete microwave systems.

Systems of this kind have been used when beacon facilities were to be added to a microwave radar system and it was not feasible to add microwave beacons to aircraft already equipped with lower-frequency beacons. The adaptation was made by adding lightweight-microwave receivers of the crystal-video type and special uhf transmitters to the beacons, which supplied the needed power. The improvisational nature of the system is evident; this did not detract from its utility. If the entire system had been designed from the beginning, a complete microwave system might have been used.

There are, however, certain advantages to crossband beacons which may recommend their use even in systems engineered from the beginning. They can be somewhat lighter than an all-microwave beacon, at least with current design techniques. Although this difference is not large and can be expected to decrease, it is still important.

More significant, however, is the possibility of providing a system capable of working with a much greater variety of interrogating equipment than is possible with any single-band system. A standardized reply frequency, which all interrogators would be equipped to receive, could be used with an interrogation band as wide as one would like—up to several thousand megacycles per second or even more, assuming this to be desirable.

*Crossband Systems with* $\lambda_i > \lambda_r$.—Because of their less directional interrogation, crossband systems with $\lambda_i > \lambda_r$ would give unwarranted triggering of beacons not being used. However, the more directional response link would permit display of signals with narrow arcs at all ranges. If both links are microwave—for example, 10-cm interrogation, 3-cm response—the objection of nondirectional interrogation disappears, and excellent azimuth discrimination can be attained. The beacon transmitter may possibly be somewhat bigger than at 10 cm (for technical reasons only, the power-output requirement being identical), and the approximate comparisons would be with systems using 10 or 3 cm on both links. A system of this type is probably less economical in requirements for transmitted power than the inverse system.

**4·7. Sweeping-frequency Systems.**—Systems have been used in which both the beacon receiver and transmitter were swept back and forth synchronously across a band of frequencies. Others have been used in which the interrogator receiver alone swept a band of frequencies (for example, AN/APS-4).

Such systems introduce scanning losses; in addition, depending upon the constants involved, they may restrict operation. One of the sweeping beacons referred to above swept a 30-Mc/sec band once every $2\frac{1}{2}$ sec with a receiver whose r-f passband was about 3 or 4 Mc/sec wide. Such a system is clearly useless with a scanning radar; the probability that the beacon will be tuned to the correct frequency at the instant the radar happens to be looking at it is too remote. Very fast (electronic) sweeping would avoid this, but would not avoid the scanning loss.

The swept receiver of the interrogator mentioned above was intended to obviate the need for AFC of the local oscillator at a fixed frequency. It did, but at the expenditure of a considerable scanning loss since it scanned 16 to 30 Mc/sec with a 2-Mc/sec bandwidth.

The scanning loss in frequency sweeping is analogous to that resulting from scanning in space, hence its name. Its value is determined by the loss in the limiting link. For a sweep of a transmitter through a frequency band of width $W$, which is being received by a receiver of bandwidth $r$ (at the signal level used), the loss is fractionally $(r/W)^{\frac{1}{2}}$. The same equation will hold for the sweep of a receiver of bandwidth $r$ (at the signal level actually used) through a band of width $W$. For the sweeping beacon mentioned above, the loss is $(\frac{3}{30})^{\frac{1}{2}} = 0.32 = 5$ db. For the sweeping interrogator receiver, the loss is $(\frac{2}{30})^{\frac{1}{2}} = 0.26 = 6$ db.

## FREQUENCY ASSIGNMENTS

**4·8. General Considerations.**—Once the frequency region or band to be used has been decided upon, the method of allocating specific frequencies must be considered. Frequency assignments of beacons are contingent upon the use for which the system is intended. It was found very early that it was desirable for the frequency of the beacon return to be different from that of the interrogator. This permits the signal to be distinguished from ground returns by detuning. Almost all beacon return signals have been off frequency (as referred to the interrogation frequency).

Stated very generally, the problem is to discover how the beacon signal can be made to appear in a readily distinguishable manner to different interrogators. The solution is that off-frequency response must be used, not only to separate beacon replies from ground returns, but also for another reason.

Experience has shown that with high-resolution radar sets it is desirable to be able to view the radar picture and the beacon returns both separately and together. Such analysis and correlation are best facilitated by the use of off-frequency response.

*Reply-frequency Coding.*—If beacon replies are to be off frequency, the simplest arrangement is to have all beacons reply on the same frequency, which is then standard for all radar sets. It may happen, however, that it is desired to obtain more data from the beacon return than are conveniently transmitted by the reply coding. Division of beacons into a few classes may then be desirable, either for the sake of more rapid recognition of the response or because the number of codes available in the reply coding is inadequate. In this case, reply-frequency coding may offer a solution.

An example of the use of reply-frequency coding arose in the operations during the land war in Europe in 1944–1945. Both the Eighth and Ninth Air Forces had ground-radar surveillance sets, with which they followed the movements of both their own and hostile planes. Both air forces had a limited supply of airborne beacons operating with their ground-radar sets. Since the functions of Eighth and Ninth Air Force aircraft were very different, it seemed desirable to be able to distinguish readily between them. The number of pulse reply codes available in the airborne beacons was small—most of them were preliminary models with no reply coding. Other beacons provided seven reply codes. In view of the importance of distinguishing the aircraft of one air force from those of the other, and the relative lack of interest in definite identification of planes of the other air force, it was decided to use different reply frequencies for the two sets of beacons.

The decision for or against the use of coding by a change of reply frequency depends on the answers to the following questions:

1. Is it desired on occasion to remove from the screen all aircraft responses except those of one of a limited number of classes?
2. Is it better to use frequency coding for this than other methods of reply coding?
3. Is it feasible to use frequency coding of the beacon reply rather than other methods of reply coding?
4. Does the reply coding have a function other than identification?

It is easier to change the number or spacing of reply pulses than to change the reply frequency. Only when the pulse coding of the reply has a function in addition to identification should frequency coding be necessary. Reply-frequency coding might also be used in airborne beacons to indicate altitude and identity, or, in ground beacons, to indicate location and weather data.

It seems desirable to restrict coding, if possible, to pulse characteristics and spacing rather than to use frequency, both for simplicity of design and for economical use of the limited spectrum space. Reply coding is discussed in detail in Chap. 5.

**4·9. Frequency Channels for Single-band Microwave Systems.**—For microwave beacons designed to operate with microwave-radar sets of one type only or of several types in the same frequency band, the pattern of frequency allocation is simple. A fixed band is assigned to the radar sets. This band is determined by two major factors—the likelihood of interference between radar sets, and the technical difficulties of maintaining a narrow band.

It is often desirable to restrict the band of frequencies to a region having a width of only $\frac{1}{2}$ or 1 per cent. This is economical of spectrum space, a consideration that has not been important at microwave frequencies in the past but is likely to become more important in the future. Restricting the band usually permits simplification or improvement of the design of the beacon, and thus lighter weight or greater sensitivity.

Too narrow a restriction of frequency is undesirable because it might increase unduly the likelihood of interference between radar sets. This interference appears in the form of unsynchronized signals, "hash" or "fruit," on the radar screen. It is particularly evident when many radar sets are operating near each other, as in naval formations. Design of the radar receivers to have low sensitivity outside of the desired pass band and a judicious spreading of frequency in the radar sets are mitigating procedures.

A factor that in the past has determined the minimum band required for a radar set is the frequency spread of the fixed-tuned magnetrons used. Such magnetrons, when tuned to a nominal frequency by cold resonance methods before sealing off, exhibit a spread of 15 to 60 Mc/sec, or even more, after being sealed off. This factor is no longer limiting, since tunable magnetrons that can be tuned to a spot frequency at will have now been developed. Accordingly, in systems in which radar sets are fairly well separated in space, narrow frequency bands and even spot-frequency operation are now technically possible.

Allowance must be made, in defining the frequency band, for a shifting of frequency of magnetrons which often occurs in operation. Such a shift may be due to variation of the reactance presented by the output circuit to the magnetron, to heating, aging, power-supply variations, and so on. All of these can be minimized by proper design of the radar set, as they have had to be in beacons to get proper operation at a fixed frequency. To obtain maximum output and simplicity, however, such frequency shifts are often tolerated.

Some of these causes of change in frequency give rise to variations proportional to the frequency, whereas other shifts are independent of frequency. The r-f bandwidth necessary for proper pulse reproduction in the beacon receiver is independent of frequency, and the absolute width of the band required to minimize interference is likewise independent of frequency since it depends only on the radar density. On the other hand, "pulling" of the magnetron, and frequency drifts due to heating, aging, and power-supply variations are all more or less proportional to the frequency.

Which of these two sets of factors is the determining one depends upon circumstances. For high frequencies and low traffic densities, the frequency-dependent variables will determine the minimum bandwidth. At 10,000 Mc/sec an over-all bandwidth of less than 15 to 20 Mc/sec would be difficult to maintain in the field using current techniques; by applying frequency-stabilization procedures it could probably be cut to about 6 to 10 Mc/sec. At higher traffic densities at any frequency, the requirement for interference-free operation will dictate a minimum bandwidth of about 20 Mc/sec. It will be understood that more precise figures must be estimated from the detailed constants and behavior of the system.

The band for interrogator operation having been determined, the response should be located at a safe distance outside the band on a clear channel (that is, one not used for any other purpose). A reasonable distance appears to be from 5 to 20 Mc/sec. Some bands and beacon frequencies used in the past are listed in Table 4·1.

TABLE 4·1.—MICROWAVE BANDS USED FOR BEACON SYSTEMS

| System | Radar band, Mc/sec | Beacon-response frequency, Mc/sec |
|---|---|---|
| Airborne radar–ground beacon.......... | 3267–3333 | 3256 |
| Airborne radar–ground beacon.......... | 9330–9420 | 9310 |
| Ground radar–airborne beacon.......... | 2700–2900 | 2907 |

These bands, especially the first two, are wider than would be required now. Each of them could now be reduced by a factor of 2. The ground-radar band was made wide to handle high traffic density.

**4·10. Fixed-channel Operation.**—Fixed-channel operation is useful when exclusive channels are required, when traffic density is low, when considerable excess power—and thus high-sensitivity narrow-band receivers—are needed for utmost reliability, whenever narrow-band receivers are used in beacons, and in nondirectional, low-frequency beacons. In the past, systems for automatic ground control of aircraft have fallen into this category wherever the control was exercised from the ground by the interrogator signal itself. Control by radio-telephone instructions to the pilot can be carried out by independent means when scanning radar sets are used. Automatic control by way of the interrogation signal generally requires a searchlighting interrogator.

The separation and distribution of such channels depend upon the selectivity of the interrogator and beacon receivers and upon their image-rejection characteristics. Receivers used in systems of this nature should have sharp slopes to their bandpass characteristics and as good image rejection as possible. Channel separation will also depend upon frequency stability.

Maximum economy of spectrum space will be achieved by the use of interrogation coding in addition to frequency channeling. Unless such coding is used, two interrogators on the same frequency may on occasion

TABLE 4·2.—FIXED-CHANNEL BEACON SYSTEMS

| System | Interrogation channels, Mc/sec | Response channels, Mc/sec | Characteristics |
|---|---|---|---|
| 1. Rebecca-Eureka; airborne interrogator, ground beacon | A 209<br>B 214<br>C 219<br>D 224<br>E 229 | 209<br>214<br>219<br>224<br>229 | Only cross-channel operation used; e.g., interrogation on A, response on B, C, D, or E. Beacon-receiver selectivity inadequate for channel separation used—7 db down at 5 Mc/sec off frequency |
| 2. Oboe, Mark II, ground interrogator, airborne beacon | A 3150<br>B 3195<br>C 3240 | A 3135<br>B 3180<br>C 3225 | Only AA, BB, CC used. The beacon receiver had an intermediate frequency of 15 Mc/sec and a bandwidth of 10 Mc/sec |
| 3. CCB-ground interrogators (SCR-584), airborne beacon (AN/APN-19) | 2890<br>2830<br>2770<br>2710 | 2907 | All beacon responses were on the same channel. Selectivity between interrogation channels was achieved with a filter cavity (Q-500) in the antenna of the crystal-video receiver of the beacon |
| 4. CCB with interrogation coding (double-pulse type), airborne beacon (AN/APN-19A) | 2890<br>2830<br>2770<br>2710 | 2907 | Four double-pulse interrogation codes were available for each interrogation channel, making a total of 16 channels |

trigger the same beacon, even if they have directional antennas. In some control systems this will not matter; in others it will be detrimental. For example, if the beacon traffic-capacity is too low to handle simultaneous interrogation by two such interrogators and still maintain a satisfactory percentage of response, or if the repetition rates of the two interrogators are crystal-controlled and hence practically identical, the control may be jeopardized by such interference because of beacon "stealing." Interrogation "jitter" or interrogation coding will take care of this situation.

The characteristics of several fixed-channel systems used in the past are given in Table 4·2.

In Systems 3 and 4, the AN/APN-19 beacon used could be adjusted for wideband operation so that any signal in the 2700- to 2900-Mc/sec band would trigger it. This adjustment was used when it was desired to respond to all scanning radars for purposes of general surveillance. A switch permitted operation with narrow-band reception by introducing a cavity filter in the receiver antenna line. This was used to exclude all interrogations except those on one of the four frequency channels above. In addition, the AN/APN-19A introduced into the circuit, when desired, a double-pulse decoder, permitting simultaneous use of any of four double-pulse channels on any of four frequencies. This gave a total of 16 interference-free channels.

System No. 1 is an example of poor design, in that the frequency channels are too close for the selectivity available.[1] For good channel separation, signals in adjacent channels should be many decibels down as compared with the desired channel. This gives a readily computed ratio of triggering range between two adjacent channels. The proposed use of the system will determine what factor is actually necessary. Horizon-range systems with some excess power will require ratios of 200 or more in range between desired and off-frequency channels (46 db) if operation at all ranges is required. Systems to work only at long range, or within a restricted spread of range, may be designed with less adjacent channel rejection.

System No. 1 also suffers from several other flaws. The identity of interrogating and reply frequencies rules out some channels because of the occurrence of radar echoes when the two frequencies are the same. Beacons on different channels may trigger each other since the receiver frequency of one beacon may be the transmitter frequency of an adjacent beacon. "Ring-around" (continuous mutual triggering) may even occur

---

[1] In practice, the system worked satisfactorily in the field, but fewer channels were available than would have been available if adequate selectivity had been provided. The designers of the system were, of course, well aware of its faults when it was designed; this was a case in which the exigencies of war took precedence over engineering judgment.

in such systems, provided the frequencies are reciprocal and the dead time of the beacon short.[1] If one beacon triggers another, the reply of the second beacon may break through in the receiver of the interrogator that tripped the first beacon. When this happens, an undesired response from the second beacon will appear, at an incorrect range that is necessarily greater than the true range.

In System No. 2 these faults do not appear. Provided the local-oscillator frequencies of the beacon receivers are identical with the beacon-transmitter frequencies, image interference will be absent. On the other hand, the system is rather prodigal of spectrum space. A proper choice of intermediate frequency is helpful here. The channel separation in this system is large because of image rejection difficulties resulting from the bandwidth of 10-Mc/sec of the beacon receiver. Had a 30-Mc/sec intermediate frequency been used, five channels could have occupied the same spectrum space, as shown in Table 4·3.

TABLE 4·3.—REVISED CHANNEL ASSIGNMENT FOR OBOE MARK II

| Interrogation, Mc/sec | Response, Mc/sec |
|---|---|
| 3160 | 3150 |
| 3180 | 3170 |
| 3200 | 3190 |
| 3220 | 3210 |
| 3240 | 3230 |

Any intermediate frequency greater than 25 Mc/sec (one-fourth the frequency band available plus one-fourth the receiver bandwidth) will permit a choice of local-oscillator frequency which makes the image frequency fall outside the band. In System No. 1, an intermediate frequency greater than about 6 Mc/sec would be adequate.[2]

**4·11. Adjacent-channel Selectivity and Pulse Shape.**—To achieve satisfactory operation with fixed channels, it is not sufficient for the beacon and interrogator receivers merely to have adequate r-f selectivity. In addition, the spectrum of the radiated pulses from each transmitter must contain insufficient power on adjacent channels to interfere with normal operation. Thus in the system just described in Table 4·3 it is necessary for satisfactory operation that signals transmitted 20 Mc/sec off the frequency in use should not trigger the beacon.

Now, if the adjacent-channel selectivity is to be, for example, 40 db, a signal on a channel adjacent to the one in use would have to be 40 db stronger than a signal on the correct channel before triggering would occur. But if there are components of such off-frequency transmissions in the received-band intensity with more than $-40$ db relative to the

---

[1] The dead time was too long for this to occur in the Eureka beacon.

[2] The receivers used in most Eureka beacons were superregenerative because of weight and space limitations.

intensity at the main frequency, it will be impossible to achieve the required selectivity.

It is well known that a square pulse of amplitude $E$, duration $d$, on a frequency $f_0$, can be analyzed by means of a Fourier integral. The analysis shows[1] that the amplitude at any frequency $f$ will be given by

$$A(f) = \frac{Ed}{2\pi} \frac{\sin \pi(f - f_0)d}{\pi(f - f_0)d}. \tag{18}$$

Either a voltage or a current pulse may be analyzed in this way. The power at any frequency $f$ will be proportional to the square of the amplitude:

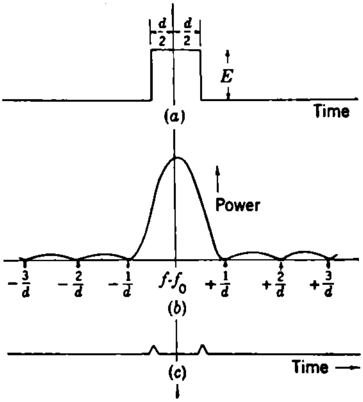


FIG. 4·1.—(*a*) A rectangular pulse. (*b*) Its frequency spectrum. (*c*) The A-scope appearance of such a pulse received off frequency.

$$P(f) = \text{constant} \cdot E^2d^2 \frac{\sin^2 \pi(f - f_0)d}{[\pi(f - f_0)d]^2}. \tag{19}$$

These relations are shown in Fig. 4·1.

TABLE 4·4.—POWER RADIATED ON FREQUENCY $f$ RELATIVE TO POWER RADIATED AT $f_0$ Expressed in db

| Pulse length, μsec | $f - f_0$, Mc/sec | | | | |
|---|---|---|---|---|---|
| | 5 | 10 | 20 | 30 | 50 |
| | db | db | db | db | db |
| 0.5 | −12 | −18 | −24 | −27.5 | −32 |
| 1.0 | −18 | −24 | −30 | −33.5 | −38 |
| 2.0 | −24 | −30 | −36 | −39.5 | −44 |

---

[1] See, for example, E. A. Guillemin, *Communication Networks*, Vol. 2, Wiley, New York, 1935, p. 463.

Let us see what the numerical values of the power radiated off frequency are. Table 4·4 gives some representative values for the maxima nearest the frequency deviation given.

It is clear that channel selectivity of 40 db is difficult to achieve if square pulses are used for transmission.

There are two ways of attacking this problem. One is to change the pulse shape in such a way that the off-frequency interference is reduced. For example, a rounded pulse with the shape of a Gauss error curve ($E = E_0e^{-at^2}$) has a spectrum with the same shape (with a change of scale), which falls off much more rapidly than the $\sin^2 x/x^2$ curve of Eq. (19). Thus, a pulse of this shape has much less power at remote frequencies than a square pulse does. The other way is to discriminate against the "spikes" produced by off-frequency components [Fig. 4·1(*c*)] by a suitable circuit in the receiver (for example, an integrator). This is equivalent to narrowing the video bandwidth of the receiver.

It is certainly more desirable to prevent the radiation of unwanted signals than to devise methods for minimizing their interference after they reach the receiver. Thus, changing the pulse shape in such a way as to eliminate or drastically reduce off-frequency components is preferable to changing the receiver.

The required change in pulse shape is of interest. Any rounding off of the steep edges of the pulse will be of value. This may be achieved in the modulator by various means. In fact, most microwave magnetrons will not perform properly with too sharp a rise of applied voltage; thus rise times (from 10 to 90 per cent voltage) of the applied pulse from the modulator seldom are less than 0.1 μsec when magnetron transmitters are used.

An even better way of minimizing the radiation of off-frequency components is to insert a tuned circuit of suitable pass band in the antenna line of the transmitter. For example, a transmission cavity with a bandwidth of 5 Mc/sec between half-power points will give the following attenuation off frequency, if the cavity is correctly matched to the line in both directions:

| Mc/sec off frequency | Attenuation, db |
|---|---|
| 5 | 7 |
| 10 | 12 |
| 20 | 18 |
| 30 | 22 |
| 50 | 26 |

The loss of power in such a cavity at the resonant frequency is given by

$$\text{loss in db} = 10 \log_{10} \frac{Q_u}{Q_u - Q_L}. \tag{20}$$

Thus, at 3000 Mc/sec, the loaded $Q$, $Q_L$, would be $\frac{3000}{5} = 600$; if the unloaded $Q$, $Q_u$, were 3000, the loss in the cavity would be $10 \log_{10} \frac{3000}{2400}$, or 1 db.

The use of a cavity filter in this way can be thought of as providing automatically just that degree of rounding off of the r-f pulse that is required to give the desired result.[1] The obvious complication of another tuning adjustment is introduced.

The pulse cannot, of course, be too drastically modified without prejudicing the precision with which range is determined or pulse width defined.

---

[1] At frequencies off resonance, the cavity will, of course, present a reactive load to the transmitter. The problems that this entails are discussed under the subject of stabilizers in Chap. 12.

# CHAPTER 5

# CODING AND COMMUNICATION

By A. Roberts

## INTERROGATION CODING

The use of interrogation coding implies the introduction into the interrogator of equipment capable of producing some special characteristic in the interrogating signal, and the introduction into the beacon of equipment capable of recognizing such a characteristic. A beacon system so equipped discriminates against signals not possessing the required characteristic. The beacon is provided with a lock for which the correct interrogation code constitutes a key.

**5·1. Functions of Interrogation Coding.**—The purposes of introducing interrogation coding into a beacon system may be any or all of the following:

1. To restrict interrogation of the beacon to a class of radar sets. An example of such restriction is the prevention of interrogation of ground beacons by airborne search-radar sets not interested in seeing them. Such restriction decreases the load on the beacon and increases its effective traffic-handling capacity.
2. To restrict interrogation of the beacon to a particular radar set. Thus, automatic-tracking sets using conical scan may sometimes require that for good tracking no other sets trigger the beacon.
3. To reduce random triggering, as, for example, that due to ignition noise or other interference.
4. To convey information to the interrogating radar set. If several different interrogating signals are available for beacon interrogation, successive trials will determine which of these will trigger the beacon. The setting of the beacon to respond to this particular signal may then convey information to the radar operator. Disadvantages of this use are obvious.

In many circumstances no special restriction need be placed on the character of the interrogating signal other than that it fall within a particular frequency region.

**5·2. Signal Characteristics.**—The beacon may utilize any of the characteristics of the interrogating signal to make its decision as to whether or not to respond. In addition to adequate intensity, these characteristics include

1. Frequency (or frequencies).
2. Number of pulses per interrogating signal.
3. Pulse spacing.
4. Pulse shape (including pulse duration).
5. Signal repetition frequency.

In practice, all of these have been used. The choice must depend on circumstances.

A decoding device should not lead to an unintentional decrease of traffic capacity. This implies that it should not introduce additional "dead" time into the system. Otherwise the presence of numerous signals of extraneous origin, which do not form the proper code pattern, will tend to paralyze the system for an appreciable fraction of the time.

**5·3. Types of Decoders: Trigger and Storage.**—Pulse decoders may be classified into two general types: trigger devices and storage devices.

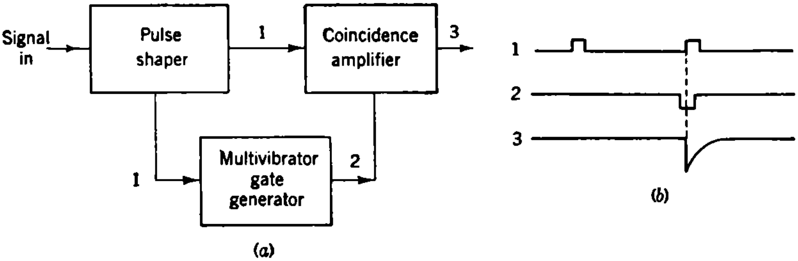


FIG. 5·1.—(a) Double-pulse decoder using multivibrator delay. (b) Waveforms at numbered points of (a).

In a trigger device, the various elements of the interrogating signal initiate a sequence of events which is carried out regardless of anything that may happen in the meantime. In a storage device, the circuit "remembers" what has happened, and acts accordingly.

As an example of a trigger device, consider a double-pulse decoder (see Fig. 5·1) which, upon receipt of a pulse, initiates a square pulse of width equal to the code spacing. The end of the square pulse generates an artificial pulse, which is fed into a coincidence circuit. If the second pulse of the code now arrives at just the right instant, it will combine with the artificial pulse in the coincidence circuit to give an output signal.

In such a circuit, every random incoming signal will trigger the multivibrator. Thus, a random pulse coming just before the arrival of a properly coded pair will prevent a response to the true code. This happens because the multivibrator circuit, being a trigger device, "shuts its eyes" after the receipt of the first pulse for the time of operation of the gate.

Let us now contrast this with the behavior of a storage device. As an example of a storage device, let us consider a circuit in which the multivibrator of Fig. 5·1 has been replaced by a delay line, which delays the original signal for a time equal to the code spacing before applying it to the coincidence circuit. The delay line is a storage circuit, since it will store, for an interval, in proper order, all incoming pulses of whatever spacing. With this type of circuit, an interfering pulse occurring just before the true code will not prevent a response to the proper code.[1]

It will be seen that storage devices are superior to trigger devices because they do not "close their eyes" but remain alert. Furthermore, in the examples given, a regular train of closely spaced pulses could effectively paralyze the trigger decoder; no matter what spacing they had, they would not affect the storage decoder.

## CODING PARAMETERS AND RANDOM INTERROGATION

**5·4. Frequency.**—The use of frequency as a coding parameter depends largely upon the number, character, and importance of extraneous signals. To begin with, any receiver defines a limited frequency range for interrogating signals under all conditions of operation.

Multiple frequencies can be used in obtaining coded interrogation signals. Beacons have been constructed which require simultaneous, or nearly simultaneous, reception of signals on two frequencies for a response to be emitted. Usually two receivers and a coincidence circuit are employed for this purpose.

Probability considerations show here that if the fraction of time during which signals arrive on frequency $F_1$ is $f_1$, and the fraction of time during which signals arrive on frequency $F_2$ is $f_2$, then the fraction of time $f$ during which signals will be received simultaneously by chance on both frequencies is

$$f = f_1f_2,$$

provided that the signals on both frequencies are independent of each other.[2] For example, if $f_1 = 0.01$ and $f_2 = 0.05$, $f = 0.0005$. In general, for $n$ frequencies, $f = f_1f_2 \cdots f_n$.

In considerations such as these the independence of the signals on one frequency from those on the other is important. If some signals are crystal-controlled in pulse-repetition frequency on both frequencies, they

---

[1] This is not always strictly true; wide pulses may cause troublesome interference even in storage-type decoders if they overlap any pulse in the code.

[2] Ignition noise has a very broad spectrum, and signals on widely different frequencies due to ignition noise are therefore not independent.

are not independent in time. Less rigorous control of the repetition rate makes for a closer approach toward true randomness.

The number of random coincidences per unit time is of more significance. If the average pulse length is $\tau_1$ on frequency $F_1$ and $\tau_2$ on frequency $F_2$, the number of pulses per unit time is $N_1 = f_1/\tau_1$ on $F_1$, and $N_2 = f_2/\tau_2$ on $F_2$. The number of coincidences per unit time is then determined by the distribution in time of $f$.

A coincidence will occur in the instances in which the beginning of one of the $N_1$ pulses on $F_1$ is followed within time $\tau_1$ by a pulse on $F_2$, or when a pulse on $F_2$ is followed within $\tau_2$ by a pulse on $F_1$. The number of such events per unit time is

$$N = N_1N_2(\tau_1 + \tau_2). \tag{1}$$

Each of the two terms represents the frequency of occurrence of these two possibilities. The time $\tau_1 + \tau_2$ is a natural or inherent resolving time that depends solely on the widths of the pulses.

The validity of Eq. (1) is subject to the restriction that both $N_1\tau_1$ and $N_2\tau_2$ are small compared with unity. To extend our example, if $N_1 = 10^4$ sec$^{-1}$, $\tau_1 = 10^{-6}$ sec, $N_2 = 10^4$ sec$^{-1}$, and $\tau_2 = 5 \cdot 10^{-6}$ sec, Eq. (1) gives $N = 600$ sec$^{-1}$. We see from this how much coincidence techniques can reduce random interrogation on frequencies on which there is a great deal of traffic.

*Resolving Time.*—We have not considered the effect of finite resolving time of the coincidence circuit. If this is long compared with the pulse widths involved, the resolving time $\tau$ is greater than $\tau_1 + \tau_2$, and the number of coincidences per unit time will be given by the formula

$$N = 2N_1N_2\tau. \tag{2}$$

Using the numerical data of the previous example, with $\tau = 5 \cdot 10^{-6}$ sec, we get from Eq. (2) $N = 1000$ sec$^{-1}$, as against 600 sec$^{-1}$ calculated from Eq. (1). Usually the resolving time can be made equal to or less than the pulse width.

A more rigorous treatment is possible but of doubtful value in practical cases. The orders of magnitude are correctly indicated by the equation given.

*Effect of Delay.*—The entire treatment above has been based on the assumption that the pulses on the two (or more) frequencies used actually overlapped in time. It is clear that none of the above considerations will be altered if a delay is inserted in the output of one receiver, and a corresponding interval between the pulses is required to give an apparent coincidence at the receiver outputs.[1]

---

[1] It must be remembered that our treatment is for *random* pulses only. Nonrandom interference, such as ignition noise, may give different results. Delayed

**5·5. Number of Pulses, Pulse Spacing.**—The use of multiple interrogating pulses on the same frequency is another way of excluding unwanted interrogations. This method places some obvious requirements on the modulator and transmitter of the interrogator. Multiple pulses can be electronically decoded by a decoder in the beacon. Timing circuits in the radar set must be triggered by the last pulse, the one which elicits the response of the beacon.

This method is often more economical of equipment than the use of multiple-radio frequencies, and involves fewer antenna problems. It is also economical of r-f spectrum space, since several pulse channels can occupy the same r-f channel.

The problem of random interrogation here takes the form: Given a system requiring, for triggering, $n$ pulses with intervening spaces $s_1$, $s_2$, . . . $s_{n-1}$, what is the likelihood that signals from different radar sets will duplicate the correct triggering arrangement by chance?

We first assume that all interfering pulses from different radar sets are independent. This is not strictly true because pulses come from timing circuits and are related by the recurrence interval. If we choose the shortest recurrence interval as our unit of time, the interfering pulses may be regarded as nearly random. In what follows, then, we are speaking of events occurring over 500- to 1000-μsec intervals. It is further assumed that the entire sequence of coding pulses is emitted in a time that is less than the chosen interval.

*Double Pulses: Storage Decoders.*—We take first the simplest case, in which two pulses must be separated by an interval between $t_1$ and $t_2$ to trigger the beacon. We assume that the decoder is of the storage type in which a pulse occurring between two pulses making up the selected interval, or just before the first one, does not interfere with the decoding action.

We now ask what is the number $a$ per unit time of random interrogations if an average of $N$ pulses per unit time are received? If $t_2$ is much less than unit time, this is given by the interval law expression[1]

$$a = N[1 - e^{-N(t_2-t_1)}]. \qquad (3)$$

This depends only on the difference $t_2 - t_1$ and not on the magnitude of $t_1$. Thus the value of code spacing is irrelevant.

If $N(t_2 - t_1) \ll 1$, as is usually the case, then

$$a \approx N^2(t_2 - t_1). \qquad (4)$$

---

coincidence between pulses at two different frequencies can, however, give improvement where none at all would be had with simple coincidence of such pulses.

[1] Derived from the Poisson distribution; see, for example, E. Marsden and T. Barratt, *Proc. Phys. Soc.* **23,** 367 (1911) and **24,** 50 (1911).

We see that the number of random triggers per unit time is comparable to that obtained from random pulses on two frequencies. In fact, for $N = 10^4$ sec$^{-1}$ and $\tau_2 - \tau_1 = 2 \cdot 10^{-6}$ sec, $a = 200$ sec$^{-1}$.

*Double Pulses: Trigger Decoders.*—We now consider decoders in which an interfering pulse arriving just before a coded gain will prevent a response. Trigger decoders are of this type. If the interval during which an extraneous pulse can prevent a response is $t_0$, the number of responses per unit time to random interrogations will be given by

$$a = N[1 - e^{-N(t_2-t_1)}]e^{-Nt_0}. \tag{5}$$

This expression differs from Eq. (3) by the factor $e^{-Nt_0}$, the probability that no pulse arrives during the interval $t_0$.

It must be noted in this connection that although a storage decoder will respond to all correctly spaced pairs (in principle) the trigger decoder will respond only to the fraction $e^{-Nt_0}$ of them.

*Multiple Pulses.*—Equations governing more complex codes containing more than one interval are readily derived from Eq. (3). For a code which contains two intervals, the first from $t_1$ to $t_2$, the second from $t_3$ to $t_4$, we obtain for the number of random interrogations of a storage decoder,

$$a = N\,[1 - e^{-N(t_2-t_1)}]\,[1 - e^{-N(t_4-t_3)}] \approx N^3(t_2 - t_1)(t_4 - t_3), \tag{6}$$

since the coincidences are independent, and so on for more intervals. Multiple-pulse coding is thus as good as coincident multiple-frequency interrogation for reducing random triggering in this general case.

Similar development of Eq. (5) gives for trigger decoders the equation

$$a = Ne^{-Nt_0}[1 - e^{-N(t_2-t_1)}][1 - e^{-N(t_4-t_3)}] \tag{7}$$

for two consecutive intervals.

**5·6. Pulse-width Coding.**—Pulse-width coding has been widely used for airborne radar sets interrogating ground beacons. It is easy to produce a wider pulse in almost any modulator, and pulse width is easy to decode. Consequently this type of coding has met with considerable favor. Pulse-width decoders introduce no dead time.

A standard system now in use requires that search pulses from airborne radar sets be shorter than 1 $\mu$sec or longer than 5 $\mu$sec. Beacon-interrogating pulses must be between 2 and 4 $\mu$sec long. This system allows both short and long pulses for search, but it preserves a distinguishable beacon interrogating pulse. Pulse-width coding can be combined with other forms of coding as desired.

**5·7. Other Interrogation Coding Systems.**—Additional tumblers can be added to the lock on a beacon as desired. It is possible to require that incoming signals have the correct pulse-repetition frequency before a response will be given. Such a pulse-repetition frequency-selection circuit generally follows other circuits that have already excluded signals other than those of proper spacing and proper radio frequency.

## OTHER INTERROGATION CODING CONSIDERATIONS

**5·8. Slow Interrogation Coding.**—We have been considering what may well be called "fast" interrogation coding, in which each interrogation signal is itself coded. Obviously "slow" interrogation coding is likewise possible. In such a system, the beacon would reply only to interrogation pulses within some time interval $t$, which might be as long as a few seconds, and which would follow the transmission of a series of keyed pulses conforming to a specified, predetermined code such as a Morse letter. The code would take a time of the order of 1 sec to transmit. Such methods have been used in automatic distress-signal systems.

The disadvantages of systems that take a long time to operate and that unlock the beacon for a long period of time are obvious. Such systems are useless with narrow-beam scanning radar sets.

**5·9. Limitations on Interrogation Coding.**—Codes may become garbled in various ways in transmission. This must be considered when reliable operation under all circumstances is a prime necessity.

If a signal can arrive at a beacon over two paths differing in length by more than the length of a pulse, a double pulse will be received. This must be considered in double-pulse systems in which the interval is to be short.

If the path difference is less than the pulse width, the pulse may be "stretched." Such stretching can be a serious annoyance in pulse-width coding. Proper siting and suitable receiver design are both remedies for this difficulty.

## REPLY CODING

**5·10. Reply Coding Parameters.**—For reply coding, the characteristics of the return signal must be arranged so as to convey information in a useful form. Let us consider these characteristics. We have at our disposal

1. Frequency of each pulse.
2. Number of pulses per reply.
3. Duration and shape of each pulse.
4. Spacing between pulses.
5. Relative amplitude of pulses.

These variables have almost all been used in existing systems.

Because the reply frequency is capable of practically infinite variation, its use as a coding parameter places certain obvious requirements on the receiver used to detect it. The number of pulses per reply has been from one to six. The duration of reply pulses has been varied over a hundredfold range; it is capable of conveying much information. Pulse shape (other than pulse width) has not been used; it requires careful design and somewhat elaborate equipment. Pulse spacing is a widely used variable and an easy one to control in transmission and reception. Relative amplitude is an inefficient parameter from a signal-to-noise standpoint.

## DISPLAY CONSIDERATIONS

**5·11. Cathode-ray Tube Displays.**—The type of coding used, of course, depends on the display.

*Deflection-modulated Displays.*—With deflection-modulated displays (for example, types A and L) and nonscanning beams, all the possibilities mentioned can be used. In practice, changing the pulse duration by Morse code has proved satisfactory in many applications. This is usually interpreted by eye; it is, therefore, called "slow" coding, since only a few Morse characters—usually about 10 to 20—are transmitted per minute. This is generally adequate for identification. Electronic reading of such characters is, of course, feasible. For example, it is possible to turn the slow code into an audible Morse signal.

*Intensity-modulated Displays.*—With intensity-modulated, persistent, map-type displays and narrow beams, a series of spaced pulses is universally used. This type of coding is called "fast" coding or "pip" coding. Spaces generally run between 5 and 40 $\mu$sec. For this type of display, amplitude modulation is not very practical. Pulse-width variation can be used, but cannot be varied in time except at a rate either slow or fast compared with the scanning rate. With accurate measurement of the interval between pips, much information can be conveyed. For visual interpretation only, it is found that two spacings, "short" and "long," are all that can be used safely unless expanded displays and calibrated time scales are available.

*Azimuth "Chopping."*—It is sometimes possible to introduce a kind of azimuth coding on PPI displays. This can best be applied in those cases in which signals occupy an appreciable angle (10° or more) on the tube. Azimuth coding is best accomplished by interrupting the interrogation at a frequency suitable for "chopping" the arc into small segments with clearly visible spaces between them. Such a display, which is not, strictly speaking, a form of response coding, is quite distinctive and readily recognizable.

When the interrogating signal is the radar signal itself, azimuth coding is not always feasible. If no special interrogation code is used, the chopping can be accomplished by suitable interruption of the response signal. If an interrogation code is used, it can be accomplished by having only a suitable fraction of pulses transmitted with the correct code for interrogating the beacon. When a separate interrogator-responsor is used in conjunction with a radar set, the interrogator trigger can be interrupted.

Interruption of the interrogation is preferable to artificial chopping of the response because it is more economical of beacon traffic capacity. Also, interruption at the interrogating end is preferable to interruption at the beacon end of the system. This is true both for the sake of economy of transmitted signals and because a fixed rate of interruption might not be suitable for all the different radar sets that would interrogate the beacon. With suitable electronic instrumentation for decoding replies, all the possibilities mentioned can be used.

One difficulty encountered with azimuth chopping is that any clutter on the screen which is interrupted by the chopping will tend to take on the appearance of coded signals. When heavy clutter is likely to be encountered, it is important not to interrupt it by the chopping method used. Such interruption of "fruit" cannot be avoided. Azimuth chopping is thus undesirable when many beacon responses are present.

**5·12. Range Coding and PPI "Clutter."**—One of the drawbacks of range or pip coding is that it takes up a considerable amount of space on the display. Six-pip codes from ground beacons take as much as 10 miles of range, and could take more if it were not for the fact that the longer codes that are possible with six pips are not often used. If many beacons are to be viewed simultaneously on a PPI, overlapping and "clutter" may assume serious proportions.

Two problems which arise in connection with PPI clutter must be carefully distinguished. One is the problem of *distinguishing* beacon signals—that is, of discerning the existence of separate responses appearing as coded groups from different beacons. The other is the question of *legibility*—the problem of providing codes that can be readily identified by eye.

*Distinguishability.*—In a complex situation, when many beacons are present, overlapping of codes can become serious. Experiments have shown that beacons spaced in azimuth by less than the beamwidth of the radar, or in range by less than the over-all code length when in the same azimuth, cannot be distinguished easily.

For a 3-cm airborne radar, this may mean that the ground beacons should be separated by perhaps 3 to 5° in azimuth, and by 10 miles in range when at the same azimuth. The 5° az separation is important at long range. At 115 miles, this requires a separation of 10 miles between beacons for good resolution. Ordinarily beacons for aid to long-range navigation are not spaced as close as 10 miles. However, even when there is confusion, the greatest course error that can occur is the beamwidth of the radar—in this case, 5°. Furthermore, the confusion will disappear and the course could be corrected when the aircraft comes nearer the beacons. For a ground radar with a 1° beam, azimuth confusion occurs if beacons are separated by less than 2 miles at a range of 115 miles. We may conclude that clutter is of little significance for navigational beacons.

Airborne beacons that happen to be at nearly the same azimuth are unlikely to remain so. Clearly, azimuth confusion is of less consequence than range confusion in general. This emphasizes the importance of using coding methods that do not take up too much range.

*Legibility.*—Increasing the number of distinguishable pulse spacings and utilizing other methods of reply coding will tend to reduce the number of pips required for the code, and thus to reduce the length of the code. Any such coding methods, however, result in codes that are not as readily identified by eye on compressed sweeps. They can be identified on expanded sweeps by eye, or decoded electronically independently of the display. Decoders that generate artificial signals can furnish no solution to the problem of putting a very large number of codes on the display because, if this were possible, the satisfactory reply codes could be emitted by the beacons.

We see then that for easy legibility on compressed sweeps a compromise between number of codes and density of beacons is necessary. The merit of the compromise is greater, the greater the number of possible codes per mile of range. No extensive experiments have been conducted to determine the number of codes legible by eye as a function of sweep expansion. Experiments that have been performed have been concerned mainly with distinguishability and have shown that beacons separated in angle by the beamwidth are readily resolved at all ranges, and beacons at the same azimuth are confused only when the codes overlap.

*Number of Distinguishable Beacons.*—These facts allow us to write an expression for the maximum number of *distinguishable* beacons on a PPI or B-scope. Let

$\theta$ = angle covered by scan (360° for PPI),
$\beta$ = arc width (equal to the nominal interrogator azimuth beamwidth),
$l$ = sweep length in miles, and
$c$ = code length in miles.

The maximum number of distinguishable beacons $N$ will then be[1]

$$N = \frac{\theta l}{\beta c}. \tag{8}$$

---

[1] It is assumed that means are provided to keep the arc width independent of range, such as time-varied gain.

As an example, for a 100-mile PPI sweep, an arc width of 5°, and a maximum code length of 10 miles, $N = 720$. For a sweep of 200 miles, an arc width of 1°, and a code length of 2 miles, $N = 36{,}000$.

The foregoing calculation is, of course, unduly optimistic, because it is based on extremely unrealistic assumptions concerning the distribution of the airborne beacons in space. The maximum number we have defined is not a practical working figure, because the relative motion of radar and beacons continually changes the configuration and a symmetric distribution is never achieved. However, the numbers are so large that even if we say that one-twentieth of the maximum is the largest figure that is safe operationally, the number of distinguishable beacons in the two cases will be 36 and 1800.

In the latter case, the problem of legibility then becomes one of obtaining a number of codes not exceeding 2 miles in length, of such a nature that at least 1800 different codes are distinguishable. As has already been intimated, no adequate experimental data on this subject exist. The possibility of achieving adequate legibility on PPI tubes certainly depends on the sweep expansion—the number of miles per inch. If we assume that the eye can distinguish readily distances differing by about 1 mm, a sweep expansion corresponding to 1 μsec/mm (2.5 miles per inch) would permit visual distinction between pulse lengths and pulse spacings differing by 1 μsec. A 12-in. PPI would then contain a 15-mile sweep, which seems adequate for dense traffic, such as that at an airport.

**5·13. Distinguishability of Beacon from Radar Signals.**—Whenever both radar and beacon signals are displayed simultaneously on a single cathode-ray tube, it is desirable to be able to distinguish one type of signal from the other at a glance. Any special characteristic of either signal which assists in doing this is valuable.

Although the various kinds of coding we have discussed aid in making this distinction, other means are available and should be used whenever possible. These include the use of "two-tone" presentation and the use of different colors for the two kinds of signal—"two-color" presentation.

*Two-tone* presentation depends on the use of different intensity levels for two kinds of signal. By using different video limit levels for the beacon and radar signals, the maximum intensity of the beacon signals can be made significantly greater than the maximum intensity of the radar signal. This simple method is effective because beacon signals are usually strong and can be made to limit easily. Good contrast is readily achieved. The system clearly will not work well for weak (not limited) signals.

*Two-color* presentation is much more spectacular, considerably more difficult, and extremely effective. It is achieved by using two display tubes giving different colors, one for each type of signal, and combining the two displays by optical methods.

**5·14. Coding by Amplitude Modulation.**—The amplitude modulation of received radar or beacon signals can be detected readily by appropriate circuits in the interrogator receiver. Cathode-ray tubes are not suitable indicators if the pulse is displayed directly on the screen, because the eye is unable to follow rapid fluctuations in intensity or amplitude though it may perceive that they are present.

Audio presentation of the modulation is possible. A very ingenious method that has been used involves applying the audio modulation to a set of tuned reeds which cover the portion of the audio spectrum desired, and noting which of the reeds are set into vibration. This is a visual display of the audio spectrum of the modulation. By this means it has been shown that most radar echoes contain appreciable modulation. The echoes from aircraft, for example, are modulated by the motion of the propellers, and the percentage of modulation is high. If the number of blades in the propeller is known, it is, in fact, readily possible to deduce the speed of revolution of the engine.

Beacon replies may be subject to similar modulation in certain cases. The reply from an airborne beacon may be modulated by the propeller if the propeller blades can intercept the radiation. More important, however, is the possibility of deliberate modulation of the beacon reply.

It will be seen that maximum effectiveness of amplitude modulation of the reply will occur in searchlighting systems. Data of this type are not suitable for transmission by scanning systems.

Experiments have shown that modulation of beacon replies well over 50 per cent can be achieved. The modulation frequency then constitutes a parameter which is, in principle, continuously variable from zero to the maximum and which can be superposed upon the pulse-repetition frequency as a carrier. Furthermore, the harmonic content can also be varied at will. The amount of information transmissible in this way is, as shown in Sec. 5·18, the amount that can be carried in a channel having a width that is half the pulse-repetition frequency.

It is usually easier to modulate a beacon transmitter in frequency than in amplitude, but more space in the spectrum is needed. An amplitude-modulation receiver of bandwidth less than twice the frequency deviation will receive a modulated signal from such a transmission. To obtain the best results, however, an auxiliary detector of frequency modulation to follow the i-f stages of the receiver should be used.

## Number of Reply Codes

**5·15. General Considerations.**—The number of different codes needed depends upon the purpose of the beacon system.

The number of navigational ground beacons seen at one time need rarely, if ever, exceed two or three for entirely adequate navigation. They should be spaced 30 to 100 miles apart. If allowance is made for repetition of identifying codes after a suitable distance, 50 seems an adequate number of codes. Two to six pips with either narrow or wide spacings are enough to provide this number.

The transmission of other information over the ground beacons must then be by means of coding of frequency, pulse duration, or amplitude. Special means must be provided for the interpretation of any of these. It is, of course, possible to transmit to the aircraft a great deal more information than is at present conveyed by the beacon-reply code.

With deflection-modulated displays, fewer beacons can be interpreted at one time than on a PPI, but the number is always sufficient for navigation.

With airborne beacons, it may be desired to obtain a great amount of data from the reply. These may include

1. The identification of the aircraft carrying the beacon—as to class or even serial number.
2. The altitude of the aircraft which is more readily determined in the aircraft than from the ground, and which is of importance for traffic control.
3. Information of miscellaneous type—distress signals, requests for information, or any other standardized and uniform data.

As this shows, airborne beacons may need to return more data than do ground beacons, and may thus require more complex coding. On the other hand, they need also to be simpler, smaller, and lighter. This dual requirement imposes a burden on the ingenuity of the designer.

Up to the present, the spacing between pips in microwave beacon responses has been used for visual identification only, generally on long sweeps. Only short and long spaces were used, centered at 15 and 35 μsec, respectively. There seems to be no objection to using spaces accurately adjusted to one of 10 values between about 10 and 30 or 5 and 25 μsec, each representing one digit. Four such digits would give 10,000 possible codes that could be electronically decoded and set up on counters. Such a code would identify an aircraft by serial number. The obvious limitation on this procedure is confusion caused by the overlapping of codes from different aircraft.

Other information, such as altitude, can use either another space or another variable. Since altitude information ought to be transmitted in such a way that aircraft changing altitude could be classified at two levels when in an overlap region, the use of a continuous variable is indicated.

Frequency, code-space, and pulse width are all available, although use of any one of them involves difficulty.

**5·16. Pip-coding.**—In pip coders generating range-coded signals, the number of possible codes varies exponentially with the number of different code spaces.

If there are $p$ code pips and $n$ distinguishable values of code spacing, the number $N_p$ of different codes with $p$ pips is

$$N_p = n^{p-1}.$$

In addition, all codes with fewer pips are possible. The total number of possible codes, then, is

$$N = \sum_p N_p = \sum_{k=1}^{p} n^{k-1}. \tag{9}$$

To get large numbers of codes it is necessary to provide a large number of pips, since the number of pips appears as the exponent in the expression for the number of codes. In the important case that $p = 6$, $n = 2$ (long and short code spaces only), $N = \sum_{k=1}^{6} 2^{k-1} = 63$.

Thus, the number of codes increases exponentially with the number of pips. The amount of coding equipment required to generate the codes increases only linearly, or even somewhat more slowly, however, because each additional pip adds one more pip generator and spacing control. Consequently, when it is not certain that the number of codes required will remain small, a small saving in coder design is very expensive in the number of possible codes and may be false economy.

The above arguments apply to any coding or decoding system in which circuit elements contribute digits in a code. Each digit represents an exponential increase in the number of codes, and a linear or slower increase in the circuit complexity.

**5·17. Combined Pulse-width and Pulse-spacing Coding.**—A coder utilizing both pulse spacing and pulse width as coding parameters has a large number of possible codes. A three-pulse coder has five variables—as many as a six-pip coder using pulse spacing only. The number of possible codes is not usually so great, however.

If there are $p$ pulses, there are $p - 1$ spaces, and $2p - 1$ total variables. If each variable may take on $n$ values, the number of possible combinations is $n^{2p-1}$. If, in addition, the number of pulses can be any number from one to $p$, the total number of possible codes is

$$N = \sum_{k=1}^{p} n^{2k-1}. \tag{10}$$

This is not so large as the corresponding number for the pulse spacing coder with $2p$ pulses, which was shown in Eq. (9) above to be $\sum_{k=1}^{2p} n^{k-1}$.

A more general expression for the case in which the pulse width may take on any one of $m$ values and the pulse spacing may take on any one of $s$ values is of interest. If there are $p$ pulses and $p - 1$ spaces, they may be arranged in $m^p \cdot s^{p-1}$ different combinations. The total number of codes possible by varying the number of pulses from one to $p$ is now

$$N = \sum_{k=1}^{p} m^k s^{k-1}. \tag{11}$$

For $p = 2$, $m = 3$, $s = 5$, $N = 48$,
For $p = 3$, $m = 3$, $s = 5$, $N = 723$,
For $p = 4$, $m = 3$, $s = 5$, $N = 10{,}848$.

## DATA TRANSMISSION AND COMMUNICATION

An interrogator-beacon system has many of the properties of a communication system. In fact, each link of the beacon system can be said to be a communication system. The very names given to the links—interrogation and response—imply communication of intelligence.

Up to the present only that amount of communication which is included in interrogation or reply coding has been considered. It is now our purpose to consider how many and what kinds of additional data may be transmitted over the links of a beacon-interrogator system.

**5·18. Channel Width and Data-handling Capacity.**—A familiar theorem of communication states[1] that the amount of information which can be transmitted over a communication channel is independent of the carrier frequency and proportional to the absolute frequency width of the channel and to the duration of the signal. The highest modulation frequency which can be transmitted is half the width of the r-f channel.[2] Thus, a broadcast station requires an r-f channel width of at least 20 kc/sec to transmit the audio frequency spectrum up to 10 kc/sec, and the width is independent of its carrier frequency. A broadcast receiver, correspondingly, will require a 20-kc/sec r-f pass band and a 10-kc/sec audio pass band to receive the modulation properly.

We have seen that the r-f channels used for beacon systems are wide. Interrogator receivers for beacon signals are seldom less than 2 Mc/sec wide, and beacon receivers are often several hundred megacycles per second wide in r-f coverage. A more relevant characteristic is, of course, the video bandwidth. This is usually 1 Mc/sec or more, in the interroga-

---

[1] Hartley's theorem; R. V. L. Hartley, *Proc. I.R.E.* **11,** 34 (1923).

[2] Special procedures like single-side-band modulation are neglected here.

tor receiver. The video band-width of a beacon receiver may be several megacycles per second, or, in light-weight systems, it may be only a few hundred kilocycles per second.

These channels are wide enough to allow a considerable amount of intelligence to be transmitted. They cannot be narrowed without sacrificing the essential purposes of the system; for example, accurate range data require sharp leading edges on pulses in both links of the system and, thus, the ability to receive frequencies of the order of the reciprocal of the pulse rise time. A flat pulse with a rise time of 0.1 $\mu$sec requires a video pass band of 2 Mc/sec or more for reasonable reproduction. The narrow video bandwidth of lightweight beacons is used at the expense of range accuracy.

*Continuous-wave and Pulse Modulation.*—Complete utilization of the capacity for data transmission of a channel requires continuous-wave operation. The channel is not completely utilized unless data are continuously transmitted to the fullest possible extent. The use of pulses immediately imposes a severe restriction on the usable bandwidth of the channel. The use of pulses may be regarded as the superposition upon a carrier frequency of a modulation envelope which has zero amplitude most of the time, and a large amplitude during the pulse intervals. For such a case the effective width of the channel, from the standpoint of data transmission, is decreased by the duty ratio of the pulses. This decrease greatly modifies the possibilities for transmission of data, since a 2-Mc/sec r-f channel, which for c-w transmission could handle a modulation band 1 Mc/sec wide, will handle only 1 kc/sec of modulation if the duty ratio is $\frac{1}{1000}$. Of course, many such low-duty-ratio pulse systems can be accommodated on the same r-f channel if means of distinguishing them are provided at the receiver; this has been done in one beacon system.

The relation between r-f bandwidth and the amount of modulation carried by a pulse system is

$$f = \frac{2m}{wp}, \tag{12}$$

where $f$ = r-f channel required, in cycles per second,
$m$ = width of modulation band required, in cycles per second,
$w$ = average pulse width, in seconds, and
$p$ = pulse repetition frequency, in cycles per second.

*Prf-limitation.*—In a radar or beacon system the pulse-repetition frequency depends upon the maximum range $R$ of the system and is limited by the velocity of light $c$, according to the relation

$$p \leqq \frac{c}{2R}. \tag{13}$$

If a higher pulse-repetition frequency is used, replies will arrive back at the receiver after a second interrogating pulse has been transmitted; and confusion in range will ensue. A system to work to 186 miles range must, therefore, have a pulse-repetition frequency less than 500 cps; a further factor of 20 per cent or so is usually allowed in order to permit the indicator sweep circuits to recover.

This limitation of $p$ is important because $p$ cannot be made high enough to allow the pulse-repetition frequency to be used as a carrier frequency for speech modulation except for the shortest distances or in cases in which not all transmitted pulses are used for ranging. To carry speech modulation, $p$ must be about 4000 cps or more. The introduction of speech modulation is possible with relatively simple techniques of frequency or phase modulation (for examples, see Chap. 11).

From Eq. (12), we see that to transmit speech information, the audio spectrum of which is generally taken as the range from 200 to 2000 cps, a duty ratio of at least 0.0018 is required with a 2-Mc/sec r-f channel. Few radar sets or beacon interrogators have as high a duty ratio as this. Smaller duty ratios can be used if wider r-f channels are used. A 2-$\mu$sec pulse, with a pulse-repetition frequency of 400 cps, theoretically can transmit speech information on an r-f channel which is 4.5 Mc/sec wide; in practice this would probably be difficult. Data less complex than speech modulation naturally require smaller bandwidths.

**5·19. Data Transmission over Systems Using Beacons.**—The above considerations are entirely general; they will be applied briefly to radar and beacon systems. In the first place, systems designed to obtain good general radar performance are inherently unsuited for communication purposes. Good radar search performance is achieved by the use of narrow beams and scanning antennas. Accordingly, the communication links between such a radar and a beacon exist only when the beacon is being triggered and received, or while the antenna is pointing at the beacon, usually only a small part of the time. As a result, the useful communication-channel width is decreased from what it would be for a searchlighting radar by a factor equal to the ratio of the scanning angle to the beamwidth; this factor may be as high as 360. Voice communication over such systems is impossible for all practical purposes. Extremely wide r-f channels would be needed and rather esoteric techniques would have to be developed—first to store information, and then to cram it all into the short periods of communication.

Even the transmission of much simpler data, like Morse coding, is difficult to achieve with systems using scanning antennas. No attempt at communication other than interrogation and reply coding has been made for such systems. In consequence, nonscanning- or searchlighting-radar sets are of most interest as far as communication is concerned.

*Interrogation and Response Links.*—Communication on the interrogation link of a beacon system is, in principle, straightforward. It need not interfere with the beacon function unless a very large amount of data is transmitted. The essential requirements are means for suitable modulation of the transmitted signal and for demodulating the signals in order to detect the transmitted intelligence. The property of the transmitted signal used to trigger the beacon may be independent of the characteristics used for data transmission. This allows the repetition-rate restrictions imposed by the velocity of light to be disregarded for data-transmission purposes. Thus, a beacon with pulse-width discrimination would reply only to those pulses of the correct width, but other, auxiliary, pulses could be used for communication (see Chap. 11).

The response link presents a different problem, because beacon replies are not sent out at will but only in response to interrogating pulses. The carrier pulses for the transmission of intelligence are, therefore, not under complete control at the beacon; and the problem of transmitting intelligence becomes more difficult. Clearly, the difficulty is greatest when many interrogators are triggering the beacon. Thus, omnidirectional interrogator-beacon systems are not so satisfactory for two-way communication purposes as are directional searchlighting systems. Of directional searchlighting systems, the most satisfactory for communication purposes will be those in which provision is made to prevent all but one interrogator from using the beacon at a time.

*Data Transmission and Coding.*—Ordinary interrogation and reply coding differ from the data transmission we have considered above in that the intelligence they convey is not contained in the signal but is prearranged, like a cipher. The range code 2-2-2 means "this is a beacon at Deer Island, Mass., located at latitude 42° 21′ 01.8″ north, longitude 70° 57′ 32.8″ west," in precisely the same fashion as the cable address "Nareco" means "National Research Council, 2101 Constitution Avenue, Washington, D.C." Such abbreviations are extremely valuable, but it is essential to realize that their "meaning" is not contained in the transmitted signal.

## PRECISE DATA

**5·20. The Use of Precise Positional Data.**—The most fundamental data obtained from a beacon are its range and azimuth. One of the special features of radar beacons, as compared with other navigational aids, is the extraordinary precision of these data. The proper utilization of such data requires special thought.

It is readily possible to measure range to a beacon with a relative precision[1] of 10 to 20 yd, at all ranges up to 250 miles. This precision

---

[1] The absolute precision may be only 20 to 40 yd. The reason for this is the variation of signal path and of signal velocity due to changes in atmospheric propaga-

can be achieved on any radio frequency used for beacons; it is accomplished by means of quartz crystal-controlled ranging units.

By means of the techniques developed for fire-control radar, azimuths may be measured to 1 mil or better. This corresponds to an azimuth error of $\frac{1}{1000}$ of the range, or 0.25 miles at 250 miles.

There are thus two techniques available for very accurate navigation. The more accurate technique makes use of range measurements from two points on the ground at accurately known locations; two such measurements made simultaneously determine position. The other method uses precisely measured range and azimuth from one point accurately located on the ground.

Ordinary PPI displays do not permit one to take advantage of this precision. In order to use the precise data, special displays are required. These may be sweeps expanded sufficiently to allow the leading edge of the pulse to be distinguished, or they may be special ranging circuits. An expansion of at least 1 in. per mile is desirable. This obviously requires a method of delaying the beginning of the sweep by an amount that is accurately known.

With a suitably expanded display, relative motions of 5 yd can be detected readily by comparison with a range mark. Automatic range-tracking circuits have been developed which will indicate range, or range minus a fixed quantity, on a meter. Such circuits can also be made to measure radial velocity, by means of differentiation of the range with respect to time. Position can be plotted on a map on an automatic plotting board with high precision to give a permanent record.

**5·21. System Delays.**—Whenever the maximum possible range precision is to be achieved, it is especially important to be able to correct accurately for the delays inevitable in the circuits of the system. Beacon and interrogator delays are considered in turn below.

*Beacon Delay.*—The delay in a beacon is inherently inconstant because the beacon is a triggered device with a variable-amplitude trigger. It always contains a circuit (generally a blocking oscillator or multivibrator) somewhere which gives a response only when the input signal is above a certain voltage level. This circuit is usually the one triggered by either the receiver output or, if there is a discriminator, by the discriminator output. Such circuits are always regenerative and have an initial feedback gain that depends upon the input signal level. They also have a finite bandwidth. Accordingly, near the threshold, there will be a region of trigger amplitude in which the delay—that is, the time required for the output signal to reach its full value—depends upon the signal strength.

---

tion and in atmospheric index of refraction. The latter is more important, depending mainly on changes in water-vapor pressure.

With proper precautions, this variation in delay can be reduced to a small value. Noise fluctuations superposed on the signal, however, will tend to widen the threshold region. The only way to avoid this difficulty is to sacrifice sensitivity by demanding a higher signal-to-noise ratio for triggering. Over-all variations in delay of not over 0.1 $\mu$sec (17 yd in range) can be achieved without undue sacrifice of sensitivity. This value can be achieved for signals that trigger the beacon at 5 db above noise. Smaller variations may require more signal relative to noise.

Aside from inherent variations of this sort, other beacon delays should be kept to a minimum. The pulse-width discriminator introduces an unavoidable delay, but this can be maintained relatively constant. Too narrow a video pass band in the receiver is bad both for over-all delay and for variation delay. Chapter 16 deals with this subject at greater length.

It should also be remembered that some present-day aircraft, and most ships, are so large that the difference in range between the location of the beacon antennas and the point at which radar reflection occurs may be appreciable. Thus, ranges being measured to the beacon antenna may differ appreciably from radar ranges, and may be more accurate.

*Interrogator Delay.*—The trigger pulses used in the interrogation part of the circuit are usually of constant amplitude. Delays due to them are therefore fixed and are normally compensated for.

There are, however, other sources of variation of delay. Variations of signal strength in the interrogator receiver can give a shift in apparent range. In addition, in systems using interrogator-responsors in association with radar equipment, delays may be introduced if the antennas of the two systems are separated physically. A displacement of 100 ft between the antennas will give a range difference of 0.2 $\mu$sec, which will be added when the antennas point in one direction and subtracted when they point in the opposite direction, and will take on intermediate values at intermediate angles.

**5·22. Methods for Using Data of High Precision.**—High-precision range measurements are made for one of two reasons. Either the location of a moving vehicle, usually an aircraft, is being determined with respect to a good map, or a map is being made.

In either case, an understanding of the nature of maps and of geodesy is important for proper utilization of accurate positional data. This subject is outside of our province. It will suffice to point out that large-scale maps, with a scale of about 1 mile per inch (for example, 1:25,000 or 1:50,000) are required for full utilization of precise data. Such maps should have inscribed upon them coordinates suitable for the measurements being made. In the case of range measurements from two ground stations, these will be lines of constant range from the stations (circles, if the earth were spherical and the maps undistorted).

Sometimes even such maps are not sufficient. Instrumentation can be devised which will permit the navigator and pilot to follow a predetermined course. The use of plotting boards, with the desired course drawn on a map, and servomechanisms designed to steer the ship or aircraft to follow the desired course, is feasible, although complex.

# CHAPTER 6

# TRAFFIC AND ENGINEERING CONSIDERATIONS

BY H. H. BAILEY AND A. ROBERTS

## TRAFFIC CONSIDERATIONS

An important operational property of any system using beacons is its traffic capacity. By this term is meant both the number of interrogators to which a beacon can respond adequately and the density of beacons which an interrogator can trigger with satisfactory results. The problem may be viewed as one of determining both the traffic capacity of the interrogation link and the traffic capacity of the response link of the system. There are several kinds of limitations which affect these two problems under different circumstances. In this chapter these effects will be described briefly.

**6·1. Saturation of the Interrogation Link.**[1]—The limitation of possible traffic by overinterrogation comes about through a reduction in the response which the beacon will give as the number of interrogators is increased. The response is defined as the ratio of the number of responses by the beacon to the number of interrogating signals which have sufficient power to trigger the beacon. The reason that beacons do not give 100 per cent response and that the response diminishes significantly with increasing numbers of interrogators is the fundamental one that beacons are provided with a blanking gate, or "dead" time, during which an interrogating signal will elicit no response. Also, the allowable duty ratio of beacon transmitters is finite and small. To put it simply, the chance that a given interrogating signal will arrive at the beacon during a dead time caused by another interrogator increases as the number of interrogators increases.

It should be clear from Sec. 2·6 that, when the response link limits the range, any reduction in the number of response signals will entail a reduction in maximum range, according to a fourth-root law. Reduction of the response to 80 per cent has a negligible effect on the attainable range, but a 50 per cent response results in a 16 per cent loss in range.

Even when saturated signals can be received at horizon ranges, there is still a limitation inherent in the finite number of pulses that are received

---

[1] Secs. 6·1 and 6·2 by H. H. Bailey.

per scan; a reasonable limit in these cases might be 20 per cent response. Per cent response is not ordinarily included in calculations of reliable range since, as will be shown, with a proper (and practicable) design the traffic that is required to produce an appreciable range reduction is larger than is commonly attained.

The above arguments apply to any type of interrogator whether on the ground, aboard ship, or in the air, and in particular to any scanning radars operating with any kind of beacon. A somewhat different phenomenon, beacon "stealing," becomes important in the special case of airborne or other mobile beacons being tracked by searchlighting ground or shipborne radar sets. It is of particular importance if the radar pulse-repetition frequencies are high and very close together, if, for example, they are crystal-controlled. Only two radars may be involved in such a case; but, if one of them "steals" a beacon away from the other, it will continue to prevent the original set from receiving signals from the beacon until their two repetition rates have run through the necessary fraction of one period of the beat that corresponds to the difference between the two pulse-repetition frequencies. This interval may be long enough for the first set to lose its target completely. The chance of this happening is good, granted an approximate intersection of the beams, if the beacon dead time is comparable to the radar repetition period. This situation is so serious that, if a high geographical concentration of such sets is contemplated, either the system should employ some kind of interrogation coding or a suitable "jitter" should be introduced into the pulse-repetition frequency. The latter expedient has been used in several beacon systems.

**6·2. Beacon Density.**—The parameter of the system that is most important in determining the permissible density of beacons is the azimuth resolution of the interrogator (as discussed in Sec. 5·12). The actual distance between beacons may be so small that one beacon might be able to trigger another. The delay inherent in all beacons, however, ensures that any such second-hand trigger will arrive after the beacon has already received the original interrogating signal. The spurious trigger, therefore, arrives during a dead time and has no effect.

A possible exception occurs when there is a long line of closely spaced beacons, of which only those at one end are triggered by the radar. In this case the rest of the string will transmit at successively later times. This condition is not easily achieved, as we shall see in Sec. 6·9, because beacons do not ordinarily have a large range for mutual tripping. It is, furthermore, an unstable arrangement that would be unsatisfactory operationally. An ultimate minimum separation would be that distance at which beacons could burn out each others' receiving crystals. These effects, however, are seldom the limiting ones.

**6·3. Saturation of the Response Link.**[1]—Saturation of the response link is approached when the number of beacon replies appearing on the display becomes large enough to spoil it. The controlling factors in such saturation are the decrease of distinguishability and legibility of the display.

The character of the display depends strongly upon the nature of the interrogator. Narrow beamwidths and fast sweeps add considerably to the traffic capacity of the response link. Thus, a system which is saturated for one interrogator may be satisfactory for another and might even be satisfactory for the same one if a faster sweep speed could be used.

The dependence of response saturation on the azimuth beamwidth of the interrogator is obvious. The narrower the beamwidth, the fewer the beacons that are triggered at one time and the more that can be distinguished on the scope. The dependence on sweep speed is likewise clear. The smaller the area displayed on the scope, the fewer the number of signals that will appear upon it.

Both these considerations are relevant to an evaluation of the importance of unsynchronized replies or "fruit." These are replies of the beacons being viewed to other interrogators. They are unsynchronized because the pulse-repetition frequencies of different interrogators are usually unsynchronized and their effect is to add to clutter on the display. Whether or not saturation of the response link occurs will depend on the number of beacons simultaneously viewed and the number of interrogations to which they are replying. The latter, in turn, depends upon the traffic capacity of the interrogation link. *It is only when many beacons capable of a high duty ratio are displayed simultaneously that saturation of the response is likely to occur.*

When saturation of the response link takes place, two general methods of improvement, other than improvement of the interrogator, are possible. One of these is gating; this is applicable only when the beacon-carrying target gives a radar echo and the radar and beacon signals are being received simultaneously. Each radar return is made to generate a short gate so that only beacon signals falling within the gate and therefore associated with the particular radar signal are allowed to appear on the screen. Aside from its limitation to ground-radar–airborne-beacon systems, this method suffers from the drawback that the beacon range is restricted to the radar range. No improvement in range by use of beacons is possible.

A better method is that known as "video integration." In this method, all the replies received after one pulse is emitted are stored, either on an uncoated cathode-ray tube face or in a supersonic delay line. They are then compared with the signals following the next pulse, in much the

---

[1] By A. Roberts.

same way as that used for moving-target indication,[1] but with a significant difference. In moving target indication, replies from alternate sweeps are subtracted from each other so that constant replies disappear. In video integration, successive replies are added and then applied to a biased tube in such a way that pulses present on a single sweep produce no output; only constant replies appear. The resolution of the system is made sufficiently low that successive pulses from a target overlap even though the target moves between sweeps; only random unsynchronized pulses disappear. This effectively removes the fruit from the screen, and, in principle at least, can even improve the signal-to-noise performance of the system. Video integration becomes impractical when the beacon response is significantly less than 100 per cent.

Saturation of the response link occurs relatively infrequently, but it is more likely to occur with nonscanning interrogators. Most microwave beacons saturate on the interrogation link before response saturation occurs; hence this trouble is more likely to be encountered with lower-frequency sets.

In the case of tracking radars, a different approach must be used. Resolving power as such loses its significance, but the beacons must be separated enough to permit the radar to lock on the desired beacon easily and then to stay locked on without interruption until a desired objective is accomplished. In general, this constitutes a rather severe restriction on the permissible number of beacons within range, the actual numbers again being determined more by the radar than by the beacon design.

**6·4. Overinterrogation Control.**[2]—Because a beacon can receive more interrogation than it can respond to safely, and because of limitations in power dissipation in the transmitter, modulator, or other circuits, some kind of overinterrogation control is required. A dead time of suitable duration following each response will prevent overinterrogation. It will be seen later that this is unfavorable for high-percentage response at traffic levels below the permissible maximum.

It is often advisable to incorporate special circuits into the beacon to achieve overinterrogation control by limiting the response to some predetermined safe value. The question arises as to how this control is to be exercised. There are two major possibilities—control by reducing the receiver sensitivity and control by increasing the dead time. The former method will discriminate against weak interrogation signals; the latter will reduce the response to all signals impartially.

The choice of the method to be used depends on the purpose of the beacon system. All microwave beacons for navigational purposes have used the second method. It is important for distant interrogators to

---

[1] *Radar System Engineering*, Vol. 1, Chap. 16; Vol. 2, Sec. 7·9, of this series.

[2] By H. H. Bailey.

receive a response, and the amount of traffic would have to be phenomenally high before any great reduction in range would occur. It is experimentally true that under conditions of maximum interrogation the range of the large microwave ground beacons is not greatly reduced.

On the other hand, for airborne beacons used for air traffic control, reduction of receiver sensitivity is the preferable method. It is better for the effective range of the beacon to be reduced than to deprive all interrogators of an adequate response; and for control purposes the nearest interrogators must be given the preference.

## THE PROBABILITY OF BEACON RESPONSE

BY H. H. BAILEY

**6·5. Statistical Analysis of the Probability of Response.**—We return now to the primary effect which limits the traffic capacity of beacons, that is, the reduction in $W$, the probability of response or per cent response, with increasing $n$, the number of interrogators. With a knowledge of both $W$ and $n$, the duty ratio of the beacon is also known. A statistical analysis will be given using as a parameter $\tau$, the ratio of the beacon dead time to the nominal interrogator repetition period.

The validity of statistical averages must be examined in every application by considering the ratio of the time required for a random reshuffling of the relative phase of the repetition periods of several interrogators to the integration or "memory" time of the display screen plus the observer (about 8 sec). We must also consider the number of pulses per thermal time constant of the dissipation-limiting components, and the number of interrogations per scan. In beacons that have been designed to date, the constants are such that observable fluctuations are superimposed on the computed averages; these fluctuations do not detract from the usefulness of the statistical results.

*Preliminary Considerations.*—A simple but incorrect approach to a specific formulation of a relationship between $W$, $n$, and $\tau$ is to state algebraically that the probability of *not* getting a response is equal to the fraction of the time that is occupied by all the dead times. That is, $1 - W = Wn\tau$, or

$$W = \frac{1}{(1 + n\tau)}. \tag{1}$$

This formula cannot be correct[1] since it gives values of $W$ less than unity even when $n = 1$. More generally, it errs in allowing all of the dead times, including the one started by any given interrogation, to affect the

---

[1] This equation is, however, strictly correct for predicting the response of a circuit with a dead time $\tau$ to *random* interrogations, the *average* number of which per unit time is $n$.

probability that the same interrogation will obtain a response. A better approximation is to say that, given $n - 1$ interrogators, the chance that an $n$th interrogator does not get a response is equal to the fraction of the time occupied by the $n - 1$ dead times. That is, $1 - W = W(n - 1)\tau$, or

$$W = \frac{1}{[1 + (n - 1)\tau]}. \tag{2}$$

This formula goes too far in correcting the error of Eq. (1), as will be shown presently. However, it is a useful approximation since it gives more accurate results than Eq. (1) for very small values of $n$, and it has the same correct asymptotic behavior.

*Calculation of the Response for Light Traffic.*—An exact expression for $W$ can be obtained through a careful analysis of the addition of the $n$th interrogating signal to a sequence containing $n - 1$ signals. Assuming, for simplicity, that the various interrogators are using the same nominal pulse-repetition frequency, the number of dead times already existing is $W_{n-1}(n - 1)$, where $W_x$ indicates the value of $W$ that obtains under conditions of simultaneous interrogation by $x$ interrogators. Therefore the probability that the $n$th interrogation does not elicit a response is $W_{n-1}\,(n - 1)\tau$. But the $n$th interrogator is just one of $n$ interrogators, all of which have the same a priori chance of receiving responses. That is, the probability just computed for the $n$th interrogator is the probability of failure of response to any one of the $n$, or $1 - W_n$.

In symbols, then, $1 - W_n = W_{n-1}(n - 1)\tau$, with the result that we have the recursion formula

$$W_{n+1} = 1 - n\tau W_n. \tag{3}$$

Given the condition that $W_1 = 1$, Eq. (3) defines $W$ for all integral values of $n$.

The solution of Eq. (3) is the polynomial

$$W_{n+1} = 1 - n\tau + n(n - 1)\tau^2 + \cdots + n!(-\tau)^n \qquad (n\tau < 1). \tag{4}$$

This expression is valid only for $n\tau < 1$. The reason for this is the breakdown of the assumption of the periodicity of the beacon response at the nominal radar pulse-repetition frequency. This breakdown will occur as soon as all the intervals between arriving signals can become less than the dead time. It is then no longer true that the dead times cover just the period $(n - 1)W_{n-1}\tau$. The pattern of the response is not obvious; the addition of an $n$th interrogating pulse may rearrange the response pattern to many other pulses, and the effects of dead times just at the end of a repetition period will change the pattern of the response in the next repetition period.

*Calculation of the Response for Heavy Traffic.*[1]—When $n\tau > 1$ an exact formula seems difficult to derive. The following derivation approaches the right result for large values of $n\tau$, and gives results which are somewhat too large for smaller values of $n\tau$. It is therefore conservative to use this expression to estimate the loading of the beacon. The formula is

$$W_n = \frac{1}{n\tau}[1 - (1 - \tau)^n] \qquad (n\tau > 1). \tag{5}$$

The derivation is as follows. Divide the repetition period into equal spaces of length $\tau$. Let there be $s$ such spaces; $s = 1/\tau$. We now apply reasoning similar to that used above, but applied to "dead" and "open" spaces instead of to interrogations. Consider $n$ interrogations and let there be on the average $r_n$ dead spaces. Now add the $(n + 1)$th interrogation. There is a chance $r_n/s$ that this added interrogation falls in a dead space, and that the number of dead spaces remains $r_n$. There is, on the other hand, a chance $(s - r_n)/s$ that the added interrogation falls in an open space and that the number of dead spaces is $r_n + 1$. The average number of dead spaces for $(n + 1)$ interrogations is therefore

$$r_{n+1} = r_n \frac{r_n}{s} + (r_n + 1)\frac{s - r_n}{s},$$

or

$$r_{n+1} = 1 + (1 - \tau)r_n.$$

Now $r_{n+1}$ is the number of responses or $(n + 1)W_{n+1}$, giving

$$(n + 1)W_{n+1} = 1 + (1 - \tau)nW_n, \tag{6}$$

which yields Eq. (5) above.

We have thus obtained equations characterizing the beacon response to sustained interrogation at low [Eq. (4)] and at high [Eq. (5)] traffic densities.

An empirical approximation that is satisfactory at all traffic densities can be obtained from inspection of Eqs. (1) and (2) by writing

$$W = \frac{1}{[1 + (n - W)\tau]}. \tag{7}$$

This formula gives values which always lie between those of Eqs. (1) and (2). It approaches Eq. (2), as it should, for low values of $n$ and converges for all values of $n$. The results are just distinguishable from values computed from the exact Eq. (3) in practical cases for which the computations have been carried out, but the errors may be considerably

---

[1] This calculation was made by S. A. Goudsmit.

larger when the dead time is very large ($\tau \geqq 0.5$). Equation (7) can be written as a quadratic in $W$ and solved for $W$, or it can be solved for $n$ and put into a form that is much more convenient for computation, namely,

$$n = \frac{1 - W}{W\tau} + W.$$

**6·6. The Effect of Duty-ratio Limitation.**—In practice, the duty ratio of a beacon is limited, usually by the maximum thermal dissipation in one or more components. When the maximum duty ratio has been attained, the response must thereafter decrease inversely as the number of interrogations. If the allowable duty ratio is less than that imposed by the dead time alone, some type of duty-ratio limitation is called for. As an example, dead time might be increased during periods of heavy traffic.

In Fig. 6·1 the beacon response $W$ is plotted against the number $n$ of searchlighting interrogators, for various values or $\tau$. Equation (3) is used up to $n\tau = 1$, and Eq. (7) is used for higher values of $n\tau$. The actual response of a beacon to a number of interrogators is found by following out the full line for the proper value of $\tau$ until the limiting duty-ratio contour (in dotted lines) is reached. The duty-ratio contours are plotted in terms of $m$, the number of interrogators required to give the maximum allowable duty ratio (assuming uniformly spaced interrogations). If the duty ratio is limited in the beacon, the duty-ratio contour must be followed from here on, to obtain the response. If it is not, the beacon will be overloaded, but the response will be greater.

It is clear that use of a short dead time and electronic duty-ratio limitation will provide a much more satisfactory response than use of the dead time which by itself would limit the duty ratio to the required value. Application of Fig. 6·1 shows, for an example picked at random, that a beacon which must be protected against a duty ratio greater than that due to five searchlighting interrogators will give much better service under light to medium interrogation with duty-ratio limitation and a $\tau$ of 0.05 than it could with the very high $\tau$ of 0.30 (or possibly 0.25) that would be required without duty-ratio limitation. Nothing is lost or sacrificed under very heavy interrogation.

All the above considerations are directly applicable to nonscanning, that is, to searchlighting interrogators.

It is also worth noting, in view of the importance of keeping the dead time short, that it is wise to use a time constant in the overinterrogation control circuit which is roughly the same as the thermal time constant of the dissipation-limiting component of the beacon. Thus the dead time will be only as long as it needs to be to protect the beacon.

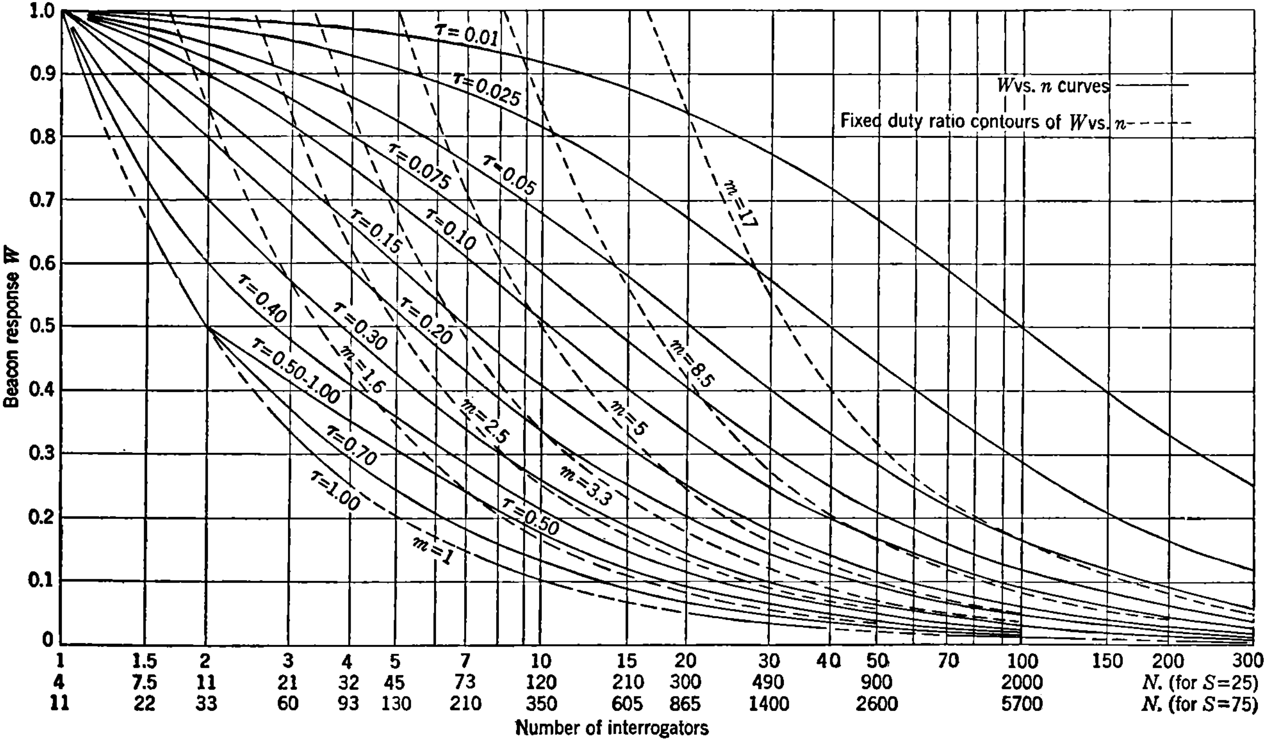


FIG. 6·1.—The per cent response $W$ of a beacon as a function of the dead time, duty ratio, and number of interrogators. Full curves show beacon response as a function of the number of interrogators for various values of $\tau$, the ratio of the dead time to the repetition period. Dotted curves show the beacon response when the duty ratio is limited so that full response can be given to a maximum pulse-repetition frequency which is $m$ times that of one interrogator. The abscissa scales show the response to $n$ searchlighting interrogators, or to $N$ scanning interrogators, which interrogate the beacon only $1/S$ of the time. For scanning interrogators the curves show the response which will be reached at least 99 per cent

**6·7. Scanning.**—It remains to superpose on these results the effects of scanning. This can be done only very roughly and with the help of several arbitrary assumptions. The complete solution to the problem of the random occurrence of related series of events is extremely difficult and complicated. However, correct orders of magnitude for the number $N$ of scanning radars that are "equivalent to" (that is, which obtain the same response and load the beacon as much as) $n$ searchlighting sets can be arrived at as follows.

Let us define a scanning factor $S$ as the ratio of the total angle scanned to the angle throughout which a beacon is triggered. Then, for averages over long enough periods of time, $N$ is obviously equal to $Sn$. But with the limited number of pulses that can ordinarily be received per scan past a beacon, such a simple result can be very misleading. The quantities that are most meaningful are the *maximum* duty ratio and the *minimum* per cent response to interrogations; hence the nature of the fluctuations from the long-term statistical averages must be examined.

We assume that the scanning period is divided into $S$ equal intervals and that the interrogations from any one radar occur entirely within one or another of these intervals. It is then a matter of straightforward computation to find the probability that any desired number of interrogations occurs within a particular interval.[1] The situation is like that of $N$ balls falling at random into $S$ slots, and the problem is to compute the chance that just $b$ balls fall into one particular slot. We further assume, quite conservatively, that the largest 1 per cent of the fluctuations can be ignored; in other words, the response of the beacon will fall below the value we calculate only 1 per cent of the time. The computation must then be extended to find the number $n$ such that the sum of the above probabilities for $1 \leqq b \leqq n$ is 99 per cent. The analogous problem with the balls is to add up the chances for $b = 1, 2, 3, \ldots$ until the total probability is 0.99; then the last value of $b$ to be used is $n$.

The result, expressed in our notation, and also in the more usual binomial form (where $p = 1/S$), is obtained when

$$\sum_{b=1}^{n} \frac{N!(S-1)^{N-b}}{(N-b)!b!S^N} = \sum_{b=1}^{n} C_b^N p^b (1-p)^{N-b} = 99 \qquad \text{per cent,}$$

which defines $N$ as a function of $n$ for any value of $S$. The calculations can be carried out by means of the normal probability integral for most values of $n$, and for small values of $n$ the binomial coefficients can be computed directly. Any value from 1 up to about 1000 can be taken

---

[1] Precisely the same result is obtained if one assumes that interrogations arrive at random instead of all within one of the $S$ intervals.

by $S$, but the results for two fairly typical values, $S = 25$ and 75, are indicated in the bottom scales of Fig. 6·1.

It must be emphasized that the scanning results are only approximate because of the arbitrary selection of 1 per cent as the fraction of the fluctuations to be ignored; there is also a difficulty in assigning a value of $S$ to any real situation because the effective beamwidth is a function of range for each interrogator. The values of $n$ can be depended upon; the values of $N$ may be in error by 30 to 50 per cent or even more, depending on many circumstances, and should be used only as an aid in estimating the real traffic capacity of beacons.

The point made near the beginning of this section, that the traffic required to limit performance significantly can be made quite large, can be demonstrated very easily with the aid of Fig. 6·1. A typical set of values for the beacon parameters might be $\tau = 0.10$, maximum duty ratio corresponding to 3.3 searchlighting interrogators. Using those curves, it is seen that some 25 sector-scanning ($S \approx 25$) radars or at least 60 medium-resolution 360°-scanning ($S \approx 75$) radars can interrogate such a beacon simultaneously with negligible loss of performance (80 per cent response). The corresponding numbers are 70 and 200 to obtain our arbitrary limit of 50 per cent response, and the numbers get up to 125 and 350 before the range is reduced by a quarter ($W = 32$ per cent).

Strictly speaking, the calculations above apply best at long ranges, where $S$ remains reasonably constant. It is of great importance to remember that at short ranges, where 30 db or more of excess power is available for interrogation, the directional properties of the interrogation may vanish completely, and that side-lobe triggering will occur at slightly greater ranges. Under such conditions it is extremely difficult to evaluate the exact amount of traffic that can be handled. The equations and curves given can be used as a guide, however, under all circumstances.

## INTERFERING SIGNALS IN BEACON SYSTEMS

BY A. ROBERTS

Like all other radio systems, beacon systems are subject to various forms of interference from unwanted radio signals. We may now consider the types of interference which are encountered, the way in which they affect beacon systems, and the measures that may be taken to counteract their effects.

**6·8. Extraneous Interference.**—Interfering signals arising outside the system may be due either to natural causes or to man-made sources of radiation. The natural sources of electromagnetic radiation in the spectrum are electrical discharges, such as lightning and corona, and thermal and spectral radiation from atoms and molecules.

Throughout most of the radio spectrum used for beacon systems, neither of these sources of interference is important. Electrical discharges—lightning, static, precipitation static, St. Elmo's fire (corona), and such phenomena—have rather broad frequency spectra, almost entirely located below 100 Mc/sec. In addition, the line-of-sight propagation characteristics of high radio frequencies result in the absence of interference from sources beyond the horizon; thus, distant thunderstorms or similar high-power disturbances do not have the long-range effects which they exercise at lower frequencies.

From the viewpoint of thermal or black-body radiation, or of characteristic spectrum lines, the radio spectrum lies in the very far infrared. It now extends so far, however, that, at frequencies above 10,000 Mc/sec, definite thermal and spectral effects of the atmosphere are observable in absorption.

The radiation spectrum of any heated body—for example, the sun—extends in theory to infinitely low and infinitely high frequencies. Most of the energy of the sun's radiation is in and near the visible spectrum, and little energy is radiated at radio frequencies. However, it is possible to detect the radiation emitted by the sun in the 30,000-Mc/sec region by radio methods, that is, with an antenna and a receiver. The radiation in the 1-cm region from a lighted match is likewise detectable at close range. The intensity of all such radiation is, however, very feeble. It requires a fair amount of experimental skill to detect it and with ordinary radar or beacon systems no perceptible interference from thermal radiation is to be expected.

The difficulties encountered from atmospheric absorption at frequencies about 15,000 Mc/sec have already been mentioned. These result from the selective absorption by constituents of the atmosphere of radiations having the same frequency as characteristic resonances of atoms or molecules in the atmosphere. It might be feared that radiation of these same frequencies could be a source of radio interference. Such interference is not observed; theoretical considerations show that its intensity is such as to render it practically unobservable.

*Man-made Interference.*—A radar or beacon receiver may pick up interfering signals by way of the antenna, in the i-f stages (of a superheterodyne receiver), in the video stages, or by way of the power supply. Pickup in i-f and video stages and from the power supply can be avoided by suitable filtering and shielding. Antenna pickup is, however, unavoidable, if the interfering signals are similar to the signals used in the system being considered.

Interference caused by radiation from other radar sets or beacons entering the antenna is minimized by improved r-f selectivity and by antenna directivity. Keeping the r-f pass band as narrow as the system requirements will allow cuts the interference to a minimum. Antenna directivity cuts down signals at the system frequencies which come from directions other than those desired. The effect of antenna directivity is measured by the ratio of the gain in the desired direction to the gain in the direction of the interfering signal; factors of hundreds or thousands are readily attained. Since microwave systems usually possess antennas of higher gain than lower frequency systems, they have an advantage in this respect.

**6·9. Interference within the Beacon System.** *Mutual Triggering.*—Beacon systems are subject to several characteristic types of interference or malfunction, most of which can be eliminated by suitable design.

Mutual triggering of beacons, or "ring-around," occurs when beacons are close enough to each other to trigger one another continuously. The range for mutual triggering is ordinarily very much smaller than the interrogation range in microwave systems because of the low gain of beacon transmitting antennas as compared to microwave interrogator antennas. With high-powered sensitive beacons working with only slight directional interrogators this problem can become more serious. The ratio of the range for mutual triggering of beacons to the interrogation range is, from the range equation,[1]

$$\frac{R_{\text{trig}}}{R_{\text{int}}} = \frac{\lambda_i}{\lambda_r}\left(\frac{P'_TG'_T}{P_TG_T}\,\frac{G'_{R\,\text{trig}}P'_{R\,\text{int}}}{G'_{R\,\text{int}}P'_{R\,\text{trig}}}\right)^{1/2}. \tag{8}$$

We have explicitly included $\lambda_i$, $\lambda_r$, and the terms $G'_{R\,\text{trig}}$, $G'_{R\,\text{int}}$, $P'_{R\,\text{int}}$, and $P'_{R\,\text{trig}}$ to emphasize that a suitably designed beacon should require very much more power $P'_{R\,\text{trig}}$ to trigger on $\lambda_r$ than the amount $P'_{R\,\text{int}}$ required on $\lambda_i$, and that the gain $G'_{R\,\text{int}}$ of the beacon-receiving antenna in the direction of the interrogator may be different from the gain $G'_{R\,\text{trig}}$ in the direction of the triggering beacon.

Mutual triggering is usually prevented by the dead time, the interval of insensitivity of the beacon for a time $t$ after it is triggered. Thus even if a beacon $A$ should be triggered by a signal, and in replying should set off a beacon $B$, the reply from $B$ cannot trigger $A$, starting a continuous ring-around, unless the time taken for the last reply pip from $B$ to reach $A$ is greater than $t$, measured from the instant $A$ was first interrogated.

An example will help to make this clear. Suppose $A$ has a dead time $t$ equal to 150 μsec and a code whose last pip is emitted 90 μsec after triggering. Suppose $B$ has the same dead time $t$ and a code whose last pip is emitted 120 μsec after triggering. Let the time for a signal to travel from $A$ to $B$ and return be $2t_{AB}$. Then, if the first reply by $A$ to an incoming signal triggers $B$, the last code pip from $B$ will arrive back

---

[1] See Chap. 2, Sec. 2·5.

at $A$ after $120 + 2t_{AB}$ μsec. If $B$ started the cycle, the last code pip would return to $B$ after $90 + 2t_{AB}$ μsec. If $2t_{AB}$ is more than 30 μsec, corresponding to a distance of about 3 statute miles, then $A$ can start a continuous ring-around (see Fig. 6·2*a*). If $2t_{AB}$ is more than 60 μsec, a trigger from $B$ can also start a continuous ring-around (see Fig. 6·2 *b*). Ring-around may occur even with a single beacon, if echoes from the surroundings return after the dead time is over. This is an infrequent occurrence.

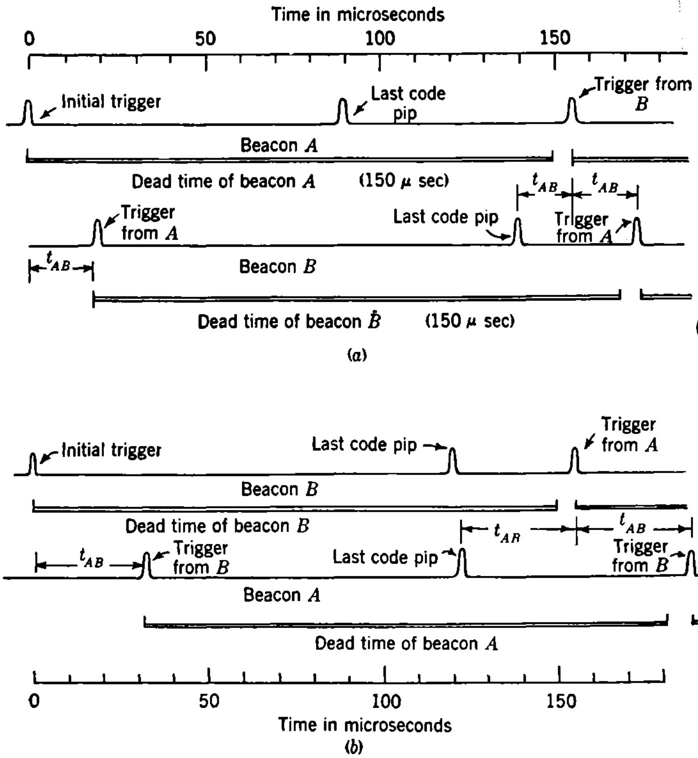


FIG. 6·2.—Mutual triggering of beacons (see text for explanation). (*a*) Mutual triggering of two beacons $A$ and $B$, initiated by $A$. The beacons are about 3 miles apart. (*b*) Mutual triggering of two beacons $A$ and $B$, initiated by $B$. The beacons are about 6 miles apart.

We see that short over-all code lengths and long dead times increase the minimum range within which ring-around is impossible. Duty-ratio limitation, which increases the dead time under conditions of high traffic density, will not stop ring-around unless the permissible duty ratio is exceeded, in which case the dead time may be increased sufficiently to stop the triggering cycle. However, it will not prevent it from starting again. Long dead times are, as we have seen, very undesirable from the standpoint of traffic capacity.

Interrogation coding of any kind greatly reduces ring-around if the beacon replies are coded differently from the interrogation. Thus, in a ground beacon responding only to interrogating pulses over 2 $\mu$sec long, the pulse-width coding will prevent interrogation by another beacon since the beacon reply pulses are only 0.5 $\mu$sec long. The interrogation coding effectively increases $P'_{R\,\text{trig}}$ by 60 db or more.

To summarize, the maximum range of mutual triggering can be minimized by decreasing the ratio $P'_{R\,\text{int}}/P'_{R\,\text{trig}}$ by means of good r-f selectivity between $\lambda_i$ and $\lambda_r$, and by interrogation coding. The *minimum* range for mutual triggering can be increased by increasing the dead time. The prevention of mutual triggering without undue sacrifice of traffic capacity then calls for keeping the dead time short, but it still calls for making the *maximum* range, as derived from Eq. (8), less than the *minimum* range, determined by the considerations of Fig. 6·2.

**6·10. Spurious Responses.**—We have seen from Fig. 6·2 that if two beacons are located sufficiently close for one to trigger the other, continuous ring-around can be prevented by a sufficiently long dead time. However, the triggering of the second beacon by the first one may not go unnoticed at the interrogator which sets off the cycle by triggering the first beacon. When the response of the second beacon is visible to the interrogator, the second beacon will appear at greater range, and possibly at an incorrect azimuth. This situation arises only when the interrogator for some reason triggers only the first beacon and not the second one. A direct path to the second beacon will always be shorter than the indirect path via the first one; thus if the second beacon replies to the direct interrogation it will be dead when the indirect interrogating pulse arrives (provided that the path difference is not so long that the time required to traverse it exceeds the dead time). Such failure of direct interrogation may occur, for example, if a hill is in the way. Since the conditions which prevent the interrogator from triggering the second beacon are likely to prevent it from seeing the spurious reply, this difficulty is rare.

Spurious responses can also appear if two interrogators have exactly the same pulse-repetition frequency. In this case the response to interrogator *A* will be synchronized on the scope of interrogator *B*, at an entirely random range depending upon the geometry of the situation and the phase of the signal. The slightest difference in pulse-repetition frequency will prevent this.

**6·11. Unsynchronized Signals.**—One type of interference which is difficult to eliminate, but which gives little trouble in general, is the appearance on the interrogator scope of unsynchronized signals. These have been discussed in Sec. 6·3. They are unmistakably different from a synchronized reply. They may even be useful if the response link is much stronger than the interrogation link; the azimuth of a beacon being interrogated by others can then be determined by the appearance of unsynchronized signals when the beacon is beyond the range for getting a synchronized reply.

**6·12. Power Considerations in the Avoidance of Interference.**—The power product theorem (Sec. 2·10) states that for systems with similar interrogation and reply frequencies and similar transmitting and receiving antenna gains

$$P_T P_R = P'_T P'_R.$$

A given interrogation range can be obtained with a high-power interrogator and an insensitive beacon receiver, or with a low-power interrogator and a more sensitive beacon receiver. Similar considerations apply to the response link.

From the standpoint of avoiding interference, it is clearly desirable to use high-power, insensitive interrogators, high-gain antennas, and high-power, insensitive beacons. The use of insensitive receivers reduces the range of interfering signals of any given power level. Freedom from interference can be bought at the price of large interrogators and large beacons.

## ENGINEERING CONSIDERATIONS

BY A. ROBERTS

**6·13. Physical Limitations and Design Economy.**—The use contemplated for a beacon often will impose limitations on the possible size, weight packaging, and power consumption, in addition to setting performance requirements. There are always limits to the minimum construction which will provide the performance required, and the task of the design engineer often is to push these limits as far as he can.

Although it is impossible to define such limits in detail in this general discussion, it will be valuable to indicate the compromises in design that involve the least sacrifice of performance for the greatest reduction in physical characteristics.

Factors tending to increase beacon size and weight are increases in

1. Pulse power output.
2. Duty ratio.
3. Receiver sensitivity.
4. Number of reply codes.
5. Number and type of interrogation codes.
6. Built-in monitoring and test features.

*Power Output.*—Decrease of modulator power output to about 2 to 5 kw pulse power is a common method of decreasing size and weight. In larger beacons, the modulator and associated power supply occupy an increasing fraction of the total size because size and weight go up sharply with increasing voltage. In one case, decreasing the pulse power output of a hard-tube modulator from 150 to 25 kw reduced the size of the modulator and associated power supply from 250 to 50 lb.

Since this decrease in power is about 8 db, it becomes apparent that improvement of the receiver of the interrogator may result in over-all economy of design. Thus, for economical design, reduction of duplexing losses to a reasonable value is essential. This has not always been achieved in the past.

When the modulator has been made as small as one of the other components, little is gained by further reduction in power output, since other factors are now controlling in over-all size.

The duty ratio of the transmitter determines the average power requirement of the modulator and its power supply; high duty ratios will increase weight, but a decrease below about 0.25 per cent gives little gain with high-power transmitters, and even larger values cost little in low-power sets.

*Receiver.*—Superheterodyne receivers are much more sensitive than crystal-video receivers. The lightest and most compact superheterodynes are comparable in size to the largest broadband crystal-video sets. For minimum size, the crystal-video receiver has not been surpassed; models the size of a cigarette case have been made.

*Interrogation Coding.*—Pulse-width discrimination imposes severe requirements on the beacon receiver, which must give faithful pulse-shape reproduction for successful discrimination. Discrimination between 1- and 2-$\mu$sec pulses demands a video bandwidth of at least 1 Mc/sec and the use of a method for achieving wide frequency coverage without sacrificing fidelity of reproduction. This rules out the use of very low intermediate frequencies. For the ultimate in light weight or portability, pulse-width discrimination must be sacrificed.

Pulse-spacing discrimination involves no such receiver restrictions, however, and may be used in lightweight and airborne equipment. Because low-power transmitters are, in general, capable of higher duty ratios than high-power transmitters, it is often feasible, in low-power beacons, to omit discrimination entirely and allow the interrogation of the beacon by all radar sets in the frequency band.

*Response Coding.*—We have seen in Sec. 5·16 that the amount of equipment required for pip generation increases only as the logarithm of the number of codes possible, or even more slowly. Accordingly, it may represent false economy to keep the number of response codes small by reducing the size of the coder, if the system is capable of using many codes.

*Monitors.*—For minimum size, monitoring and test features must be sacrificed; built-in signal generators, video amplifiers, and cathode-ray tube circuits are the first to go. Minor features—such as test points for observing voltages and waveforms, standard frequency cavities, audio amplifiers, meters, and so on—can often be retained.

*General.*—In considering over-all design it must be taken into account that for a given (low) pulse power output, uhf transmitters may be somewhat lighter than microwave transmitters. As the frequency increases, radiationless design becomes more and more imperative for triode transmitters; completely enclosed cavities which have more weight and bulk than transmission-line oscillators are required. Microwave magnetrons need magnets of appreciable bulk and weight.

This advantage of uhf transmitters is probably a temporary one which will change as techniques improve. In principle, one can expect that microwave transmitters, with smaller resonant circuits, will eventually require as little or less weight and volume than lower frequency transmitters of comparable output. This is already true at pulse power levels of 5 kw and up[1].

**6·14. Need for Attendance.** *Ground Beacons.*—When a beacon is operating properly there is usually nothing for the beacon operator to do. The beacon attendant, if any, is more properly called a maintenance man than an operator. His function is to keep the beacon in adjustment. The operation of the beacon is essentially automatic.

In the design of the beacon those features which require occasional adjustment for proper operation must be considered in the light of the amount of attention the beacon will receive. Experience has shown that well-designed beacons on all frequencies can operate with very little attention indeed. Cases of satisfactory operation for 1000 hours with no attention at all are on record.

The features which determine how much attention will be required are the following.

1. Life of components.
2. Stability of receiver adjustment.
3. Need for changing codes.
4. Stability of decoder or discriminator adjustment.
5. Stability of frequency of the transmitter.
6. Weatherproofing.
7. Independence of prime power fluctuations.

---

[1] In 1945 an experimental 3-cm magnetron (BM-50) with a pulse power output of 50 watts and weight of less than 1 lb including the magnet had been designed and was in process of redesign for production.

The mean life of the beacon between breakdowns will be determined by the component life. Good engineering practice will dictate the design and choice of suitable components.

The stability of receiver adjustment will depend upon the type of receiver used; the crystal-video receiver is probably the least troublesome. In superheterodynes, local-oscillator frequency stability is a consideration of importance. Automatic frequency control is desirable for unattended operation.

Circuits can be devised for changing codes by remote control.

The transmitter frequency is usually the most troublesome of all parameters to maintain; good design is of the utmost importance. With AFC, however, operation without attendance becomes feasible. Even with cavity stabilization alone, the frequency can be maintained within suitable limits. The bandwidth of the radar receiver used for beacon signals determines how much stability is required.

Weatherproofing is an obvious necessity for proper operation. This applies particularly to antennas and r-f transmission lines. Independence from fluctuations in line voltage, and the like, is desirable for unattended operation and implies the use of well-stabilized power supplies.

*Airborne Beacons.*—Operators are not usually assigned to airborne beacons. Any adjustment required, such as change of code, either will be made by a crew member who has other duties or will be performed automatically.

The applicable design considerations are similar to those for ground beacons. Servicing and adjustment will be performed on the ground; the interval between servicings need not be as long as it is for ground beacons. The conditions of operation—temperature, pressure, and so on—will vary more widely than on the ground, and power sources may be less constant and reliable.

Beacons designed for operation without continuous attendance often incorporate remote monitors and alarms. Thus, the interrogation of the beacon can be monitored by audio means from a distance, and a qualitative idea can be obtained of how well the beacon is operating. Remote alarms can give warning of power failure, blown fuses, open relays, or other failures.

Permanent records of beacon operation are likewise possible. Recording milliammeters or watt-hour meters can give a record of the time and intensity of interrogation.

## PART II

# BEACON DESIGN

In the following chapters the components of beacons and their integration into complete beacons are described. Most of the components are described from the point of view of the system's engineer rather than from the viewpoint of the component designer. Although this treatment results in what may appear to be a disproportionate emphasis on certain components, it is adopted in order to avoid duplication of detailed descriptions given in other volumes of the Radiation Laboratory Series. References are given to such descriptions. It is intended that enough material be included in all cases to give a coherent, self-sufficient treatment.

# CHAPTER 7

# BEACON DESIGN: R-F COMPONENTS

BY W. A. DOWNES, E. R. GAERTTNER, S. E. GOLIAN,
J. J. G. McCUE, AND T. S. SAAD

## BEACON ANTENNAS[1]

Because beacons do not have to scan and usually must have wideband receivers, it is often easier to provide them with two separate antennas than to design a broadband duplexing circuit to use with a single antenna. Because the received power is much smaller than the transmitted power, the power-handling capacities of the two are different. Their required bandwidths are also, in general, different. In the following paragraphs, requirements common to all types of beacon antenna are discussed.

### GENERAL REQUIREMENTS

The parameters that must be considered in the design of any beacon antenna are pattern, gain, frequency, and polarization. Because a beacon must necessarily operate in conjunction with an interrogator, these parameters for its antennas are determined largely by the system requirements. Other factors that must be considered are power-handling capacity and case of construction. The way in which each of these factors affects the design is discussed in the following sections.

**7·1. Gain.**—Antenna gain is a measure of the concentration of power radiated in a particular direction. Since there is a fixed amount of power to be divided over all space, it is evident that high gain is always obtained at the expense of angular coverage. One of the basic problems of beacon-antenna design is to make the best compromise between gain and coverage. In general, the beacon antenna must be omnidirectional in azimuth to allow interrogation and reply in all directions. As a result the maximum gain is relatively small compared with that of a parabolic reflector, which concentrates the radiated power in azimuth as well as elevation.

The maximum value of gain that is of practical use increases somewhat with frequency. At the lower frequencies the maximum attainable gain for a practical antenna may be limited by its physical size. Thus, a 100-Mc/sec beacon antenna of gain 10 would be 15 m high. At micro-

---

[1] By E. R. Gaerttner and S. E. Golian, with contributions from C. L. Longmire.

wave frequencies maximum gain is limited not so much by physical size as by the required coverage. Values of maximum gain greater than about 20 usually give coverage in elevation that is inadequate for use with airborne interrogators. The gain requirements for various types of beacon installation are discussed in Sec. 7·6.

*Frequency.*—In general, a beacon is required to receive over the band of frequencies available to the interrogator-transmitter, and to reply on one or more selected frequencies. This means that the receiving antenna should be matched over the whole required operating band. The antenna for the transmitter needs to be matched only for the beacon reply frequencies. In practice, it is often advisable to make the bandwidth of antennas reasonably wide and to use auxiliary devices, such as tuned r-f filters, for any r-f selectivity that may be necessary.

**7·2. Antenna Patterns.**—The antenna pattern is a three-dimensional plot of the relative power radiated in all directions. The "free-space" pattern is determined theoretically by the nature of the radiator. In practice, however, the pattern is affected by reflection, interference, diffraction, refraction, and absorption of the radiated power. The actual pattern depends on the beacon site. It is assumed in the following paragraphs, for the sake of discussion, that the beacon site is ideal—that is, that the free-space patterns are approximately realized.

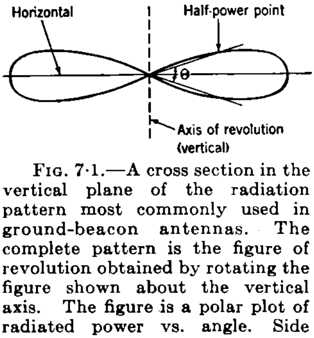


FIG. 7·1.—A cross section in the vertical plane of the radiation pattern most commonly used in ground-beacon antennas. The complete pattern is the figure of revolution obtained by rotating the figure shown about the vertical axis. The figure is a polar plot of radiated power vs. angle. Side lobes are omitted.

The pattern most commonly used is the one shown in Fig. 7·1. It is a figure of revolution about a vertical axis through the beacon antenna; consequently, it is omnidirectional in azimuth. The elevation pattern is determined by the vertical half-power width, which is usually given in terms of the angle Θ enclosed by two vectors drawn to the half-power points as shown in Fig. 7·1.

**7·3. Coverage.**—By coverage is meant that region in space within which the beacon can be interrogated and the interrogator can receive a response. It is clear that the coverage depends on the patterns of both beacon and interrogator antennas, as well as on other parameters, of which the most important are transmitter power and receiver sensitivity. The way in which the coverage is influenced by the patterns of the interrogator and the beacon is illustrated in Fig. 7·2.

Figure 7·2*a* illustrates the relationship between an airborne interrogator and a ground beacon at long range. Because of the large distance between the two, the maximum gain of each antenna is effective. With sufficient transmitter power and receiver sensitivity in each link, the maximum range may then be limited by the horizon only.

The relationship between interrogator and beacon at short range is shown in Fig. 7·2*b*. It is clear that for an aircraft flying toward the beacon at a constant altitude, a point is eventually reached at which the product of the gains measured along the line joining the two becomes too small to give the interrogator a response. The point at which this happens depends, of course, on the altitude, the tilt of the interrogator beam, and other parameters of the beacon and radar. It can be calculated from the range formula[1] if all factors are known. The region within which the interrogator can no longer receive a response because of failure of one or both of the links is called the "cone of silence."

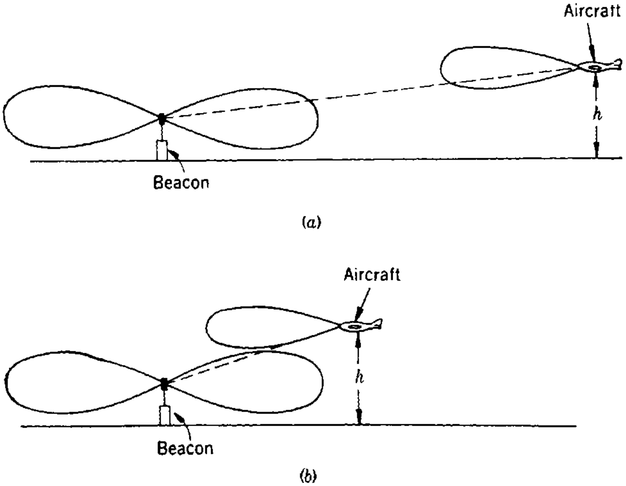


FIG. 7·2.—The relationship between the antenna patterns of an airborne interrogator and a ground beacon. The curvature of the earth is not shown. The maximum gains are effective at long range, but the useful gain diminishes at shorter ranges.

In Fig. 7·3 is shown a plot of the relative power received by the beacon or the interrogator as a function of range and altitude for a number of values of tilt of the interrogator antenna. It is assumed that both antenna patterns are like those shown in Fig. 7·2.

If the maximum gain of the beacon antenna is in the horizontal direction, and the pattern is symmetrical, as it usually is, half the power of the beacon will be radiated downward. Much of the power of a ground beacon will be wasted unless the beacon is at an unusually high elevation, with a sky line free of obstructions. In order to improve the

---

[1] See Chap. 2, Sec. 2·5 of this volume.

coverage by directing more of the radiated power into the sky, the beam may be tilted upward slightly. The surface of revolution described by the vector denoting maximum gain is then no longer a plane but an inverted cone of half angle somewhat less than 90°. Not much tilt can be permitted, since maximum gain is still desirable in directions close to horizontal. A suitable compromise seems to be a tilt of about one-quarter the beamwidth; this reduces the gain at the horizon about 1 db (see Fig. 7·4).

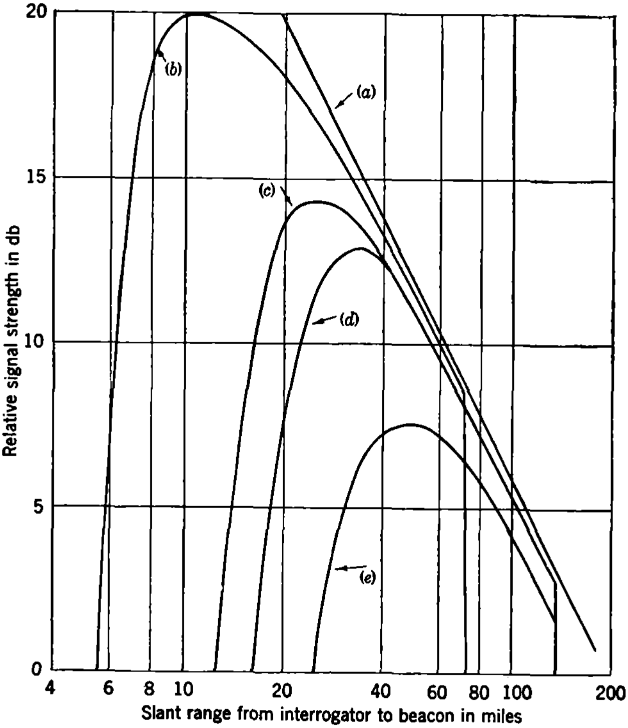


FIG. 7·3.—The strength of the signal received by a ground beacon from an airborne interrogator under various conditions. It is assumed that the airborne interrogator antenna has an elevation half-power width of 6°, and that the beacon receiving antenna has a half-power width of 10°, with the axis of the beam horizontal. (*a*) Inverse square signal dependence. (*b*) Received signal for interrogator altitude 0.6 miles, 0° tilt of beam. (*c*) Interrogator altitude 2.4 miles, antenna directed at beacon. This shows the effect of the beacon antenna elevation pattern. (*d*) Interrogator altitude 2.4 miles, antenna tilt 3° below horizontal. (*e*) Interrogator altitude 2.4 miles, antenna tilt 0°.

*Overhead Coverage.*—If coverage above a ground beacon is a prime objective, it can be approximated by using a cosecant-squared pattern for the interrogator antenna and an isotropic pattern for the beacon antenna, as illustrated in Fig. 7·5. In a cosecant-squared pattern, the gain within the antenna lobe is proportional to the square of the cosecant of the angle $\theta$ measured from the horizontal plane through the aircraft. With this arrangement, both antennas have appreciable gain near the vertical direction. This system provides not only overhead coverage but also a strength of reply that is nearly independent of range. This is evident from the following consideration: for a constant beacon-antenna gain $G'$, an interrogator gain $G$ proportional to $\csc^2 \theta$, and a constant altitude $h$, the factor $G'G/R^2$ appearing in the range equation is independent of the slant range $R$, because $\csc^2 \theta = R^2/h^2$.

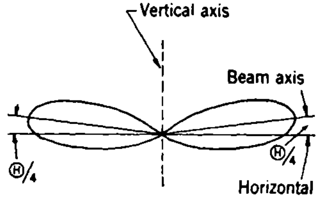


FIG. 7·4.—Antenna pattern with axis of the beam elevated above the horizon by one-fourth the beamwidth.

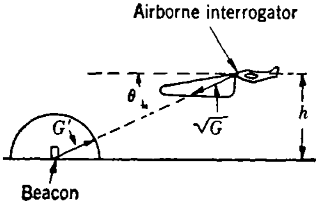


FIG. 7·5.—Overhead coverage with isotropic pattern of the beacon antenna and cosecant-squared pattern of the interrogator antenna.

A disadvantage of this type of beacon antenna is that the gain at low angles may be too small to give satisfactory range. It is possible, however, to make an antenna with a pattern that is effectively a combination to give both range and coverage overhead as shown in Fig. 7·6. The gain at low angles for such a pattern is, of course, less than that of the standard pattern because considerable power is used in its isotropic portion. For example, if the low-angle gain in the standard pattern is 10, then for added overhead coverage of gain 1 the low-angle gain is reduced to about 5.

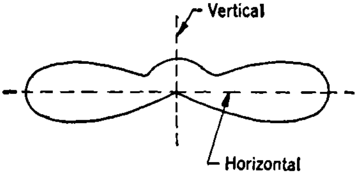


FIG. 7·6.—Antenna pattern of a ground beacon that combines the standard pattern with overhead coverage. Maximum gain at low angles has been decreased to provide overhead coverage.

## BEACON ANTENNA DESIGN

The design of antennas varies greatly with frequency over the range of frequencies from 100 to 30,000 Mc/sec. Over this range, the length of a half-wave dipole varies from 150 to 0.5 cm. In the following paragraphs, the design of a number of antennas for different microwave frequency ranges is discussed. The discussion will be limited to antennas that are omnidirectional in azimuth.[1]

**7·4. Microwave Antennas.** *Ten-centimeter Antennas.*—In the neighborhood of 3000 Mc/sec, the basic element used for a horizontally polarized omnidirectional antenna consists of three dipoles mounted in a circle as shown in Fig. 7·7. It is often called a "tripole." This combination gives an azimuth pattern uniform to within 3 db in power. When several of these elements are combined, the resultant pattern is considerably more uniform. The dipoles are excited by means of pins that extend into the coaxial line. The gain for a single element is about $\frac{3}{2}$ in the plane of the dipoles. Higher gains can be obtained by stacking the elements coaxially to form a linear array as shown in Fig. 7·8. In

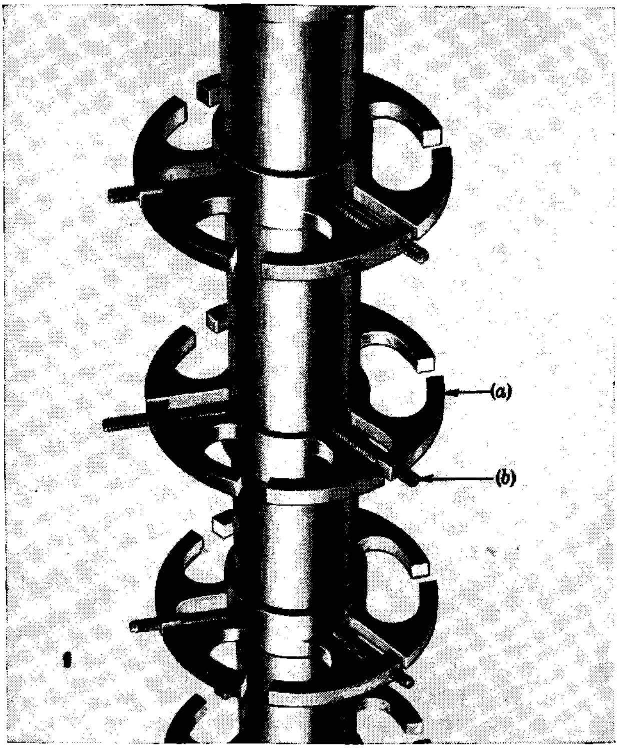


FIG. 7·7.—"Tripole" (horizontally polarized) linear-array antenna for use at 3000 Mc/sec. Note phase reversal of alternate elements. (a) Tripole element. (b) Exciting screw.

---

[1] For a complete discussion see *Microwave Antenna Theory and Design*, Vol. 12, Chap. 9, Radiation Laboratory Series.

the linear array, the elements are spaced one-half wavelength apart along the coaxial line, and they are caused to radiate in phase by reversing the phase of alternate elements. The pattern of such an array is similar to that shown in Fig. 7·1.

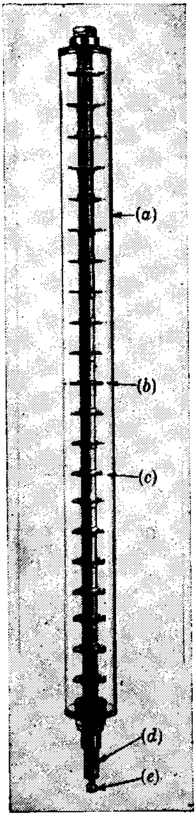


Fig. 7·8.

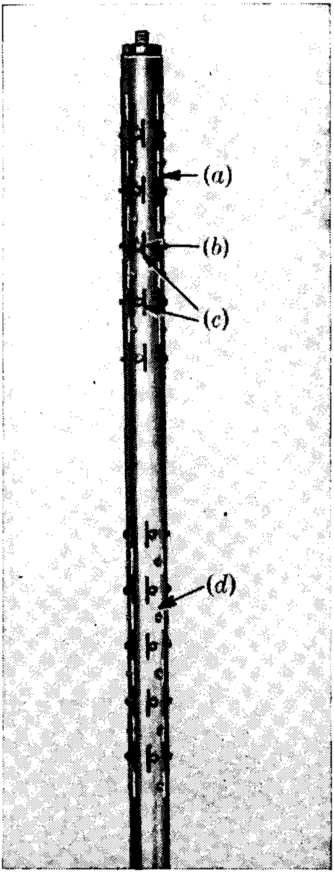


Fig. 7·9.

Fig. 7·8.—Horizontally polarized, coaxially fed, "tripole" linear-array antenna for 10-cm beacon, complete with pressure-tight housing. It includes two 10-element arrays, one for transmitting and one for receiving. (*a*) Receiving antenna. (*b*) Isolating diaphragm. (*c*) Transmitting antenna. (*d*) Transmitting antenna input. (*e*) Receiving antenna output.

Fig. 7·9.—Horizontally polarized, coaxially fed, linear-array, "leaky pipe" antenna for 3-cm use. The pressure-tight housing (similar to that shown in Fig. 7·8) is not shown. There are two 10-element arrays, one for transmitting and one for receiving. Note the phase reversal of alternate elements. (*a*) Receiving array. (*b*) Radiating slot. (*c*) Exciting screws. (*d*) Transmitting array.

*Three-centimeter Antennas.*—At about 10,000 Mc/sec and above, dipole elements usually are not used because of their small size. At these frequencies a more convenient radiator is the so-called "leaky pipe" (slotted coaxial line or waveguide) shown in Fig. 7·9. It consists of a series of narrow slots in the wall of the transmission line with pins extending into the transmission line adjacent to the slots for extracting power. The slots are cut parallel to the r-f line and are about one-half wavelength long. It can be shown that such a slot behaves like an electric dipole rotated 90°. A ring of slots in a plane perpendicular to the transmission line (four slots in Fig. 7·9) is analogous to the three-dipole 10-cm element discussed above. A 3-cm linear array consists of a series of such rings with their centers separated by one-half the wavelength in the transmission line, with phase reversal on alternate elements.

*Gain and Beamwidth.*—For all types of linear array considered here the maximum gain $G$ is approximately equal to the number of half wavelengths contained in the length of the array.

Thus

$$G \approx \frac{2d}{\lambda}, \tag{1}$$

where $d$ is the length of the array and $\lambda$ is the wavelength measured in air. The beamwidth $\Theta$ at half power is given approximately by the equation

$$\sin\frac{\Theta}{2} \approx \frac{\lambda}{2d}. \tag{2}$$

The relation between gain and beamwidth is therefore

$$G \approx \csc\frac{\Theta}{2}. \tag{3}$$

The gain and beamwidth for a given number of elements depend on whether the feed is a coaxial line or a circular waveguide. This is true because the spacing of the elements is determined by the wavelength inside the r-f line, which is greater in waveguide than in coaxial line. Hence, for a given number of elements, a waveguide array is longer than a coaxial-line array. Equations (1) and (2) are always applicable.

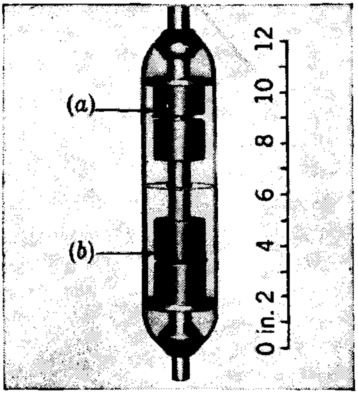


FIG. 7·10.—Vertically polarized antenna for use at 3000 Mc/sec. (a) Transmitting antenna. (b) Receiving antenna.

In a vertically polarized array suitable for microwave frequencies, the elements radiate parallel to the feed. At 3000 Mc/sec the radiator is a cylinder. Because the radiator is symmetrical about the vertical axis, a pattern is formed which is omnidirectional in azimuth. Arrays are constructed of the cylindrical elements in the manner described above. An example is shown in Fig. 7·10. Vertically polarized beacon antennas for use at 3 cm have not been designed.

*Antennas for Overhead Coverage.*—In addition to the arrays discussed above, which, in general, give patterns of the type shown in Fig. 7·1 or Fig. 7·4, arrays can be made to shape the beam into other types of pattern. Patterns suitable for overhead coverage, for example, can be made in this way by adjusting the separation of two or more adjacent elements in the linear array to less than one-half wavelength, thus adding some "end fire" radiation to the standard pattern. Patterns with uniform gain between about 20 and 70° from the vertical direction have been obtained in this manner. The decrease in low-angle gain will, of course, depend upon the percentage of the power used in overhead coverage.

**7·5. Constructional Features.**—Dual antenna systems may be made by mounting the receiving array directly above the transmitting array. This arrangement is necessary for high-power systems to prevent coupling between the transmitting antenna and the receiving antenna that would damage the receiver crystal. The upper antenna may be fed by a line that passes in front of the lower antenna. The effect of this line on the lower antenna can be minimized by placing the line at an angle to its axis and at some distance from it.

To eliminate the external transmission line that feeds the upper antenna, double coaxially fed antennas may be used. Examples of such antennas for 3-cm and 10-cm operation are shown in Figs. 7·8 and 7·9. In this construction, the lower antenna is fed by a coaxial line which uses, for its inner conductor, the outer conductor of the coaxial line feeding the upper antenna. Because of its simple shape, this type of antenna is useful in airborne and portable installations. In both types of construction, however, precaution must be taken to prevent coupling of the transmitting and receiving antennas.

*Power-handling Capabilities.*—The power-handling capability of any antenna is limited by voltage breakdown. For the 10-cm three-dipole radiator, the voltage breakdown occurs between the probes and the inner conductor. For the 3-cm leaky pipe, the breakdown occurs across the slots. Obviously trouble from breakdown may be expected to be greater for short arrays, in which the power per element is greatest. The measured value of breakdown power for a two-element, leaky-pipe, circular waveguide array is about 50 kw. A single tripole element with its probes 0.5 mm from the center conductor will not break down at 10 kw.

**7·6. Beacon-antenna Requirements and the Nature of Beacon Installations.**—The design of a beacon antenna depends on the installation for which it is intended, whether for a permanent, portable, or airborne installation. If the beacon is a permanent ground or shipborne installation, physical size and weight of the antenna are secondary to reliability. But if the beacon is to be highly portable, antenna size and weight become important. An antenna designed for extreme portability must be built of the lightest possible material. If the beacon is to be airborne, the antenna must be lightweight and also streamlined to reduce aerodynamic drag. Problems peculiar to permanent, portable, and airborne installations, respectively, are considered in the following paragraphs.

*Ground Installations.*—Permanent ground installations are usually put on hills, tall buildings, and the like, to get the least departure from free-space conditions. The gain may be as large as is consistent with the desired coverage. A small beamwidth is no handicap because the antenna can be vertically aligned as accurately as desired.

*Portable Beacon Installations.*—The site for a portable beacon cannot always be selected with much care. It may be in fields, in woods, and in other areas where natural obstructions are encountered. Further, because of the haste with which most portable installations are made, the antenna may not be accurately aligned vertically. For these reasons, it may be desirable to use antennas of lower gain than those used for the permanent ground installations. In portable installations, the antenna shape is more important than it is in a ground installation. In microwave beacons requiring two separate antennas, coaxially fed antennas of the type shown in Figs. 7·8 and 7·9 are the most serviceable.

*Ship Installations.*—A ground antenna can be kept vertical, but a shipborne antenna, unless vertically stabilized, pitches and rolls with the ship. If the antenna beam is narrow, the consequent motion of the pattern may interfere seriously with consistent beacon operation. Two alternatives suggest themselves: vertical stabilization of the antenna or large antenna beamwidth. Since a beamwidth of 30° or more is needed to counteract the motion of the ship, a serious loss in gain may result. This loss in gain is particularly unfortunate at microwave frequencies, because of the large transmission-line losses likely to be present in most ship installations; the shipborne beacon and its antenna often must be widely separated. For example, the antenna may be mounted on the mast and the beacon be installed below deck. Accordingly, if the system is one with small margins of power, gyrostabilization of shipboard antennas is advisable when possible.

*Airborne Installations.*—The motion of an aircraft in flight makes it more difficult to get proper coverage by an antenna. Furthermore, obvious aerodynamic requirements and the possibility of screening by portions of the aircraft increase the difficulty. Ideal coverage would be such that satisfactory operation with the interrogator could be obtained to maximum range irrespective of the aspect of the aircraft. Thus, an ideal condition calls for either a spherical pattern or limitation of the aspects for which the system is required to operate.

Although free-space conditions can be realized to a fair degree in most ground installations, this is generally not true for aircraft because the antenna is necessarily mounted near the skin of the aircraft, which is usually metal. As a result, the antenna pattern is usually affected by interference between the radiation from the antenna and from its image in the skin. The resulting pattern must usually be determined by measurement, although it can be computed in some instances.

Although it has been found desirable to place the antenna some distance from the skin of the aircraft to reduce such interference effects and to get clear of obstructions, it is difficult to do this without either greatly increasing the aerodynamic drag or sacrificing the mechanical strength of the antenna support. Nevertheless, some installations have been made on aircraft fuselages with supports allowing the antenna to be placed some 6 to 10 in. from the aircraft skin. An installation of this type makes the beacon pattern less dependent on the configuration of the particular aircraft in question (see Sec. 20·8 of this volume).

Duplexed antennas, double coaxially fed antennas, and separate receiving and transmitting antennas are all used in aircraft installations. The duplexed system is preferred because of its small size. It is useful at ultrahigh frequencies, but has had limited use at microwave frequencies because of the relatively high transmitter power currently required for broadband microwave duplexed systems. Double coaxial feed permits one antenna to be mounted directly above the other; such a combination is often small and easily streamlined.

If separate antennas are used, they must be mounted in such a way as to give a minimum of screening, and far enough apart to ensure that the transmitted power will not damage the receiver crystal. The minimum required separation $d$ can be conservatively calculated from the range equation, assuming that the burnout power of the crystal is known. The equation is

$$d = \frac{\lambda_r}{4\pi}\sqrt{\frac{P'_T G'_T G'_R}{P_B}},$$

where $\lambda_r$ is the wavelength, $P'_T$ is the transmitter power in peak watts, $G'_T$ and $G'_R$ are the gains of the transmitting and receiving antennas, respectively, and $P_B$ is the pulse power in watts at the receiving antenna terminals required to damage the receiver crystal. The r-f line losses have not been specifically included, but they must, of course, be taken into account. For example, with $\lambda_r = 10$ cm, $G'_T = G'_R = \frac{3}{2}$, $P'_T = 100$ watts, $P_B = 1$ watt, and negligible r-f line losses, the separation should be at least 30 cm.

## COMPONENTS FOR R-F TRANSMISSION

In the following sections some of the r-f components which lie between the transmitter-receiver and the antenna will be discussed very briefly. They include transmission lines, duplexers, r-f filters, and certain r-f accessories. Certain other r-f components, namely, superheterodyne mixers, crystal detectors for video receivers, and magnetron stabilizers will be discussed in connection with the appropriate receiver and transmitter components. Others, such as transitions, joints, couplings, waveguide bends, twists, and so on, will not be discussed at all; they are not peculiarly related to the beacon art.[1]

**7·7. Transmission Lines.**[2]—Over the range of frequencies from 100 to 30,000 Mc/sec the choice of a suitable transmission line[3] varies considerably. At microwave frequencies, the useful transmission lines are coaxial lines and waveguides. At lower frequencies, coaxial lines and two-wire lines are used.

*Transmission at* 3 *Cm.*—At 3 cm, r-f power is usually transmitted through air-filled rectangular waveguides. Two sizes are in use: 1 by ½ in. by 0.050-in. wall, and 1¼ by ⅝ in. by 0.064-in. wall (outside dimensions). The larger guide is used in installations requiring very long lines such as ship installations and permanent ground installations because the attentuation of r-f power is less in the larger guide than in the smaller guide. The main advantages of the small guide are its size and its weight, which is about one-half as much per unit length as for the larger size.

At 3 cm, air-filled coaxial line is used only in short lengths because of its high attenuation. Solid dielectric transmission lines are, in general, ruled out at 3 cm because of prohibitive r-f losses. In Table 7·1 the attenuation per meter for copper waveguide and coaxial line is given. For other metals the attenuation is proportional to the square root of the resistivity. Special materials, such as silver-plated molded plastics, can also be used when very light weight is a requirement, and in sections of guide that must provide thermal insulation. An air-filled rubber-

TABLE 7·1.—ATTENUATION BY TRANSMISSION LINES AT 3.2 CM

| Type of line | Loss in db per meter at 20°C |
|---|---|
| ½ in. air dielectric coaxial line (copper) | 0.23 |
| 1 × ½ in. waveguide (copper) | 0.117 |
| 1¼ × ⅝ in. waveguide (copper) | 0.072 |

---

[1] A complete description of microwave r-f components is to be found in *Principles of Microwave Circuits*, Vol. 8, and *Microwave Transmission Circuits*, Vol. 9, of this series.

[2] By E. R. Gaerttner, W. A. Downes, and T. S. Saad.

[3] Transmission lines are discussed in detail in *Microwave Transmission Circuits*, Vol. 9, Radiation Laboratory Series.

covered flexible waveguide is also used where a flexible link in the r-f line is desired, as between a shock-mounted receiver or transmitter and the line to the antenna. It may also be used to allow for the thermal expansion and contraction of solid lines.

*Transmission at* 10 *Cm.*—At 10 cm, r-f power is transmitted through either air-filled waveguide or coaxial lines. The size of the waveguide is $1\frac{1}{2}$ by 3 in. by 0.08-in. wall. Waveguide is less universally used than it is at 3 cm; it is used principally in applications in which very long lines or high pulse power are required. Air-filled coaxial line is generally used at 10 cm. The standard 50-ohm air-filled coaxial line has an outside diameter of $\frac{7}{8}$ in. and a stub-supported inner conductor. It has an attenuation greater than that of waveguide. Solid dielectric-filled coaxial line in the form of flexible cables is also used at 10 cm. Although lossy, it can be used for short lines. Flexible cable finds its main use in highly portable or airborne beacons in which the line lengths do not exceed about 6 to 10 ft. It may also be used where a short flexible link in a line is necessary, for example, between a shock-mounted airborne beacon and its antenna. In Table 7·2 the attenuation for waveguide and some types of coaxial line is given.

TABLE 7·2.—ATTENUATION BY TRANSMISSION LINES AT 10 CM

| Type of line | Loss in db per meter at 20°C |
|---|---|
| $1\frac{1}{2} \times 3$ in. waveguide (copper) | 0.0199 |
| $1\frac{5}{8}$ in. stub-supported coaxial (copper) | 0.038 |
| $\frac{7}{8}$ in. stub-supported coaxial (copper), RG-44/U | 0.075 |
| Polyethylene-filled coaxial cable, RG-17/U | 0.328 |
| Polyethylene-filled coaxial cable, RG-9/U | 0.558 |

*Transmission at Lower Frequencies.*—At lower frequencies, dielectric-filled flexible cables of the RG-9/U or RG-58/U type are generally used. Their use at longer wavelengths is permissible even for long transmission lines because the r-f losses decrease markedly with wavelength.

*Pressurizing.*—Moisture, in the form of vapor, has little effect on r-f transmission. If considerable changes in temperature take place, however, moisture may condense as dew inside the transmission line. Water has a very high dielectric constant and is lossy. Condensed water in a transmission line is disastrous to system performance.

Accordingly, it is essential that steps be taken to exclude water from all air-dielectric transmission lines used in field installations. This is accomplished by pressurizing the lines. The entire transmission-line system is made gastight by suitable design. The line is then kept filled with dry air at a pressure slightly above atmospheric, so that moist atmospheric air is excluded. If slight leaks are present, the pressure of dry air can be maintained by occasional pumping. In permanent installations, automatic pumps and desiccators are convenient for this purpose.

## DUPLEXERS

BY E. R. GAERTTNER AND W. A. DOWNES

A duplexer is a device that makes possible the use of a single antenna and transmission line for both receiving and transmitting purposes. It must also, of course, prevent the receiver from being damaged by the transmitter. A good duplexer must transmit almost all the transmitter power toward the antenna and almost all the received power from the antenna to the receiver.[1]

It is evident that a radar duplexer must contain nonlinear elements, since it must behave differently at high power levels (transmitting condition) than at low power levels (receiving condition). These nonlinear elements must be so situated in the circuit as to introduce impedance changes appropriate to effect the desired power transfer. They may consist of gas-filled switch tubes. Diodes have been used, as has the nonlinear grid-cathode input impedance of the first receiver tube.

**7·8. General Considerations.**—Most microwave duplexers depend for their operation on the use of gas-filled switch tubes. These are tubes filled with a suitable gas mixture at a low pressure; they contain electrodes between which a discharge can be maintained, if sufficient voltage is applied. In a duplexer, one or more such tubes are placed in the r-f circuit in such a way as to ensure that

1. When the transmitter is operating, the receiver is disconnected and all (or almost all) the transmitted power is directed to the antenna.
2. When the transmitter is inoperative, the receiver is connected to the antenna and the transmitter is disconnected so that all (or almost all) the received power enters the receiver.

These switching actions are obtained by the impedance changes provided by the firing of the gas-filled tubes when the transmitter is turned on. Switch tubes that protect the receiver are called "TR tubes," and those used to disconnect the transmitter are commonly called "ATR tubes."

An essential part of the duplexing arrangement is the use of sections of transmission line as transformers. Figure 7·11 shows the principle of the usual microwave duplexer, drawn for convenience as a two-wire transmission line.

---

[1] Duplexers are discussed in detail in *Microwave Duplexers*, Vol. 14, Radiation Laboratory Series.

1. Transmitting condition: Both TR and ATR tubes fire. Both present short circuits $\lambda/4$ from transmission line, and thus have little effect. No power goes to receiver; almost all goes toward the antenna.
2. Receiving condition: Both TR and ATR tubes quiescent, unfired. The ATR line presents a short circuit at the end of a $\lambda/2$ line, thus a short circuit at junction *a*. This short circuit appears at the receiver-line junction *b*, $\lambda/4$ away as an open circuit. No received power is then transmitted down the line toward the transmitter; all goes toward the receiver.

In the usual microwave duplexer, somewhat different r-f elements are used. The most conspicuous difference is that the TR and ATR tubes are usually in resonant cavities.

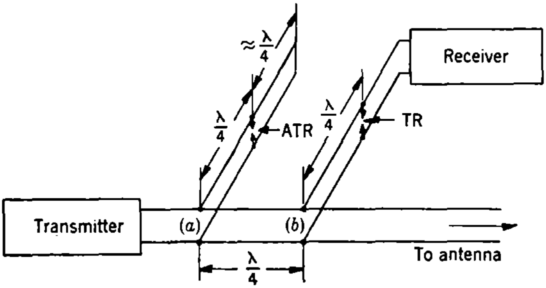


FIG. 7·11.—The equivalent two-wire transmission-line circuit of the usual microwave duplexer.

The ATR tube is sometimes omitted; this is possible if the line length from the receiver T-junction *b* to the transmitter can be adjusted so that the transmitter presents a high impedance at the junction when "cold," that is, not transmitting. This requires individual adjustment of lengths of lines for each transmitting tube if the cold impedance of the tube is not uniform enough to allow the use of a standard length of line. In the past, some types of magnetron have been sufficiently uniform; others have not.

At lower frequencies duplexing has been accomplished both by gas-filled switch tubes and by the use of grid current in the first receiver stage without gas-filled tubes. The second method is possible when the receiver uses a tube (not a crystal) as its first stage, and the average transmitter power is not large enough to damage this tube by the amount of grid current drawn. The latter restriction usually means that not more than a few watts average power can be duplexed in this way.

**7·9. Radar and Beacon Duplexers.**—In the usual radar duplexer, the TR tube is located in a resonant cavity, the loaded $Q$ of which is usually 150 to 300. This acts as a broadly tuned r-f filter in receiving; it is not nearly broad enough, however, for use in most microwave beacons, where a 1 to 2 per cent frequency band must be transmitted. For beacon use, cavities with much lower $Q$ are required. Such duplexing systems have now been designed for both 3-cm and 10-cm operation. Figure 7·12 shows a 3-cm beacon duplexer.

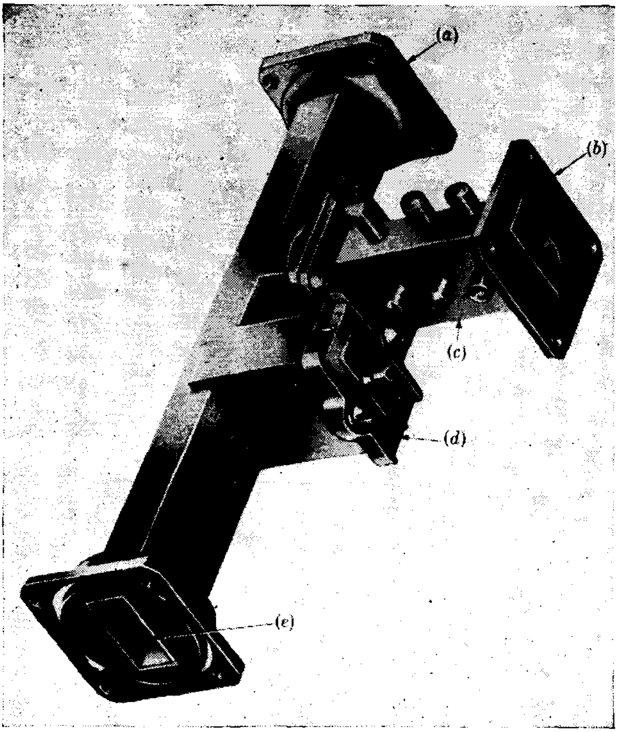


FIG. 7·12.—Broadband 3-cm beacon duplexer. (*a*) Connection to antenna. (*b*) Connection to receiver. (*c*) TR tube. (*d*) ATR switch. (*e*) Connection to transmitter.

At 10 cm either a combination of TR and ATR tubes or a special form of triode duplexer may be used. Figure 7·13 illustrates a 10-cm duplexer.

These broadband duplexers have certain disadvantages that restrict their use at present to high-power beacons. First, a minimum transmitter pulse power of 4000 watts at 3 cm and 10,000 watts at 10 cm is required to fire the low-$Q$ gas-filled TR tube.[1] Second, there is a loss in

---

[1] So far, no attempts have been made to lower these power requirements by using an auxiliary firing pulse.

received power amounting to about 2 db at the edge of the scatter bands. This loss, although small from the standpoint of range performance, nevertheless may be the deciding factor in choosing between the use of a duplexer or separate antennas.

**7·10. Duplexing without Gas-filled Tubes.**—When the input circuit goes to the grid of a tube, duplexing may be accomplished by using the change of input impedance when grid current flows. Since the transmitter and receiver are connected to the same antenna, enough of the transmitted signal arrives at the receiver to cause grid current to flow in the first input stage; as a result, there is a change in the receiver input impedance during the transmitting period. At the transmitter, there is a change in the internal impedance of the transmitting tubes as the signal is sent out. These two changes in impedance are made use of in the following duplexing schemes.

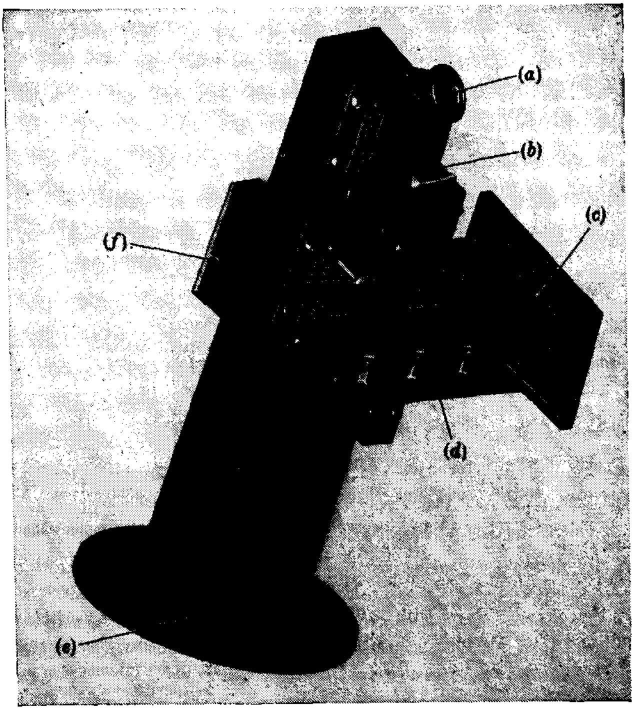


FIG. 7·13.—Broadband 10-cm beacon duplexer. (a) Connection to transmitter. Note use of two ATR switches. (b) ATR switch. (c) Connection to receiver. (d) TR switch. (e) Connection to antenna line. (f) ATR switch.

*Ten-centimeter Triode Duplexer-mixer.*—Some triodes, such as the 2C40 lighthouse tube, may be used to duplex 10-cm beacons. In this application the tube is used instead of a crystal as the mixer for a superheterodyne receiver.[1] In one type of 10-cm duplexer-mixer, the tube is mounted in the center of a flat radial cavity as shown in Fig. 7·14. Coaxial-line connections are made to the antenna and transmitter. Duplexing is accomplished in the following manner. During the transmitting period, the triode grid draws current, and, therefore, its grid-cathode impedance is low. The low impedance is converted to a high impedance at the T-junction by placing the cathode an odd number of quarter wavelengths from the T-junction. During the receiving period, the relatively high impedance of the inactive magnetron transmitter is transformed to a high impedance at the T-junction by making the distance between the two an integral number of half wavelengths. This distance may be fixed, or, if necessary, adjusted with a phase shifter. The loss in received power does not exceed 2 db anywhere within the 66-Mc/sec band, and the loss in transmitted power is about 1 db.

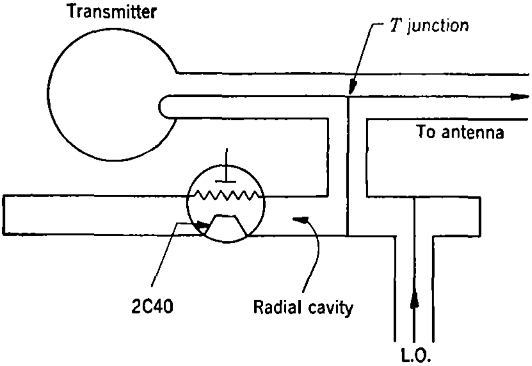


FIG. 7·14.—Ten-cm duplexer-mixer. Advantage is taken of grid current flowing during the transmitting period to present a high impedance to the transmitter signal at the T-junction.

*Ultrahigh-frequency Duplexers.*—A simple form of fixed-tuned duplexer, employed in an airborne transponder, is shown in Fig. 7·15*a*. It consists of a $\lambda/4$ section of coaxial cable connecting the transmitter and receiver. The transmitter is a push-pull pair of 6C4 tubes in a self-excited oscillator circuit, and it delivers a pulse output power of approximately 25 watts. The receiver is a special form of superregenerative detector using a type 9002 triode.

The transmitter coupling loop is adjusted to deliver maximum power output. Inasmuch as weight and space requirements are strict, no attempt is made to convert the balanced transmitter and receiver lines to unbalanced lines before duplexing.

---

[1] The performance of the tube as a mixer is discussed in Sec. 8·15.

In Fig. 7·15*b* is shown the approximate equivalent circuit for the transmitting condition, where all tuned circuits are resonant at the same frequency, and $T_t$ = the transmitter tank, $T_d$ = the $\lambda/4$ transformer, $R_a$ = antenna load, and $T_R$ = the receiver tank. The components to the right of $R_a$ present a high impedance and absorb little power.

Figure 7·15*c* is the approximate equivalent circuit for the receiving condition. The transmitter has a high impedance and absorbs little power. No data are available on the losses in this duplexing system; the losses are, however, small when transmitting and not serious when receiving.

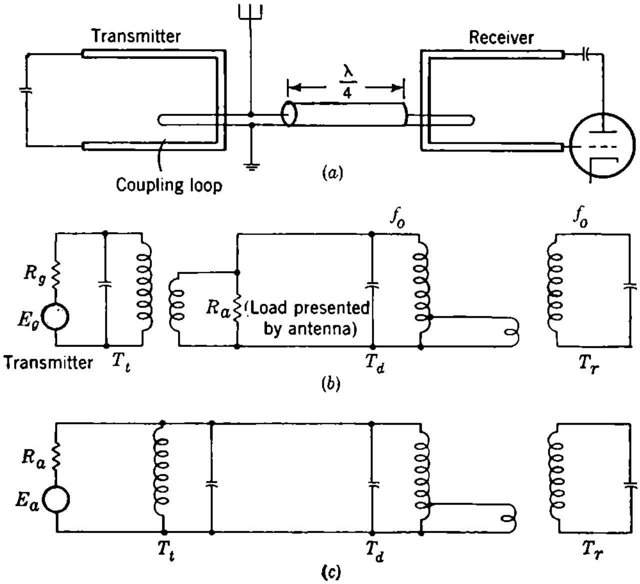


FIG. 7·15.—Equivalent circuits of a uhf duplexer. (*a*) Schematic circuit. (*b*) Equivalent circuit when transmitting. (c) Equivalent circuit when receiving.

*Duplexing with R-f Filter Protection.*—In case the receiver and the transmitter operate at different frequencies, duplexing can be accomplished without the use of nonlinear elements in the circuit. It is sometimes possible to duplex the system by placing an r-f bandpass filter between the receiver and the junction of the receiver and the transmitter line. The success of such a system depends upon the frequency separation of the transmitter and receiver, and the required band coverage of the receiver. For a "spot" receiver frequency, a narrow bandpass filter can be used successfully if the transmitter and receiver frequencies are widely separated. For the common case of wideband coverage, the design characteristics of a suitable r-f filter are difficult to meet. At 10 cm, for instance, the filter must pass a 2 per cent band without appreciable loss in received signal over the band and reject the transmitter frequency that lies only 0.3 per cent from the edge of the band. When r-f filter duplexing is used, loss of the received signal into the transmitter line is prevented by placing the transmitter at such a distance from the junction of the lines that a high impedance is presented by the transmitter line to the junction.

## Other R-f Components

**7·11. Radio-frequency Filters.**[1]—In this section we consider the use of r-f components to achieve a desired degree of selectivity. Used in this way, they may be classified as filters. As filters, in turn, they may be classified as low-pass, high-pass, or bandpass, or, similarly, as rejection filters.[2]

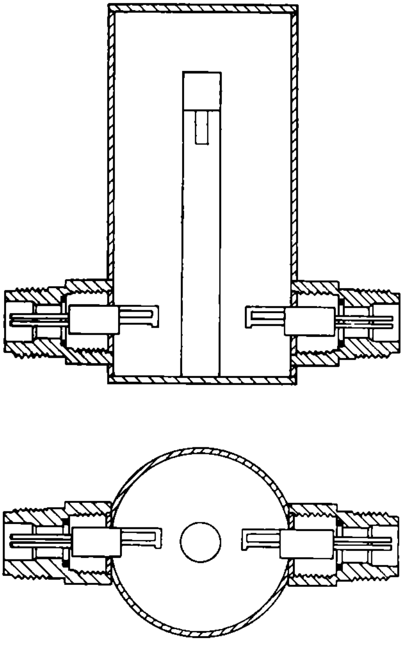


Fig. 7·16.—Single-tuned 10-cm r-f filter. The insertion loss and selectivity may be varied by rotating the coupling loops.

*Bandpass Filters.*—Bandpass filters are used to restrict the signals that reach the receiver to those occurring within a specified band of frequencies. This may be necessary to protect receiver crystals from near-by high-power radars operating at frequencies other than beacon frequencies or to prevent unnecessary triggering by signals outside the beacon band.

*Single-tuned Filter.*—In Fig. 7·16 is shown a single-tuned 10-cm filter that is used for frequency channeling in the AN/APN-19 beacon described in Sec. 4·10. It is a silver-plated coaxial resonator. The length of the invar center conductor is approximately three-fourths the resonant wavelength. By adjusting its length, the cavity is tunable from 9.0 to 11.2 cm. The loaded $Q$ is 450 to 650 for an insertion loss at the band center of 1 db, and 700 to 1000 for a loss of

---

[1] By E. R. Gaerttner and W. A. Downes.

[2] A complete discussion is to be found in *Microwave Transmission Circuits*, Vol. 9, Chaps. 9 and 10, Radiation Laboratory Series.

2 to 3 db. The band center is stable to within 1 Mc/sec for a 100°C change in ambient temperature.

*Double-tuned Bandpass Filters.*—A double-tuned filter is, in general, preferable to a single-tuned filter because its response can be flat over a wide range of frequencies. Such filters can be made by coupling two resonant circuits tuned to slightly different frequencies.[1]

At 3 cm such coupled cavities are conveniently made from sections of waveguide used as cavities. At 10 cm two cavities like that shown in Fig. 7·16 may be coupled by means of a short section of transmission line. The advantages to be secured by the use of a double-tuned circuit are better off-frequency rejection and a wider pass band with less critical tuning. On the other hand, the insertion loss is approximately doubled and the alignment of the filter is more complicated.

*Coupled-mode Filter.*—Another type of filter uses a "two-in-one" cavity, shown in Fig. 7·17. The normal $TE_{1,0}$-mode in the rectangular waveguide applies power into the cylindrical cavity through a $\frac{3}{8}$-in.-diameter circular window, exciting the dominant $TE_{1,1}$-mode. The electric vector in the cavity is excited parallel to the electric vector in the incoming waveguide. A mode transformer consisting of a screw $C$, inserted at 135° to the axis of the entrant waveguide, excites a $TE_{1,1}$-mode with the electric vector perpendicular to the original electric vector in the cavity. The power from this induced mode is then applied to the rectangular waveguide perpendicular to the original guide. The single cavity acts as two coupled resonant cavities. The set screws $A$ and $B$ are used to vary the frequencies of the respective modes in order to allow for differing input and output impedances. The pass band may be made nearly flat for a frequency range of 45 Mc/sec at 3000 Mc/sec with an insertion loss of about 1 db.

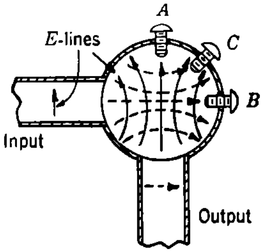


FIG. 7·17.—Ten-cm coupled-mode filter cavity.

*High-pass Filters.*—A short section of rectangular guide of proper width can be used as a high-frequency high-pass filter. For wavelengths $\lambda < 2a$, where $a$ is the guide dimension perpendicular to the electric vector in the $TE_{0,1}$-mode, the attenuation in copper guide is shown in Fig. 7·18. It is seen to be low and relatively constant for $2a/\lambda > 1.1$. For $\lambda > 2a$, the attenuation is very large; it is plotted in Fig. 7·19 in decibels per wavelength of guide. For example, a 6-in. length of guide for which $a = 1.95$ in., $b = 0.870$ in., has an attenuation of 0.01 db at $\lambda = 9$ cm

---

[1] *Microwave Transmission Circuits*, Vol. 9, Chaps. 9 and 10, Radiation Laboratory Series.

and 28 db at 10.5 cm. The filter section may be inserted in either a coaxial or a waveguide line by means of suitable couplings.

Cutoff filters of this type are useful to keep interfering signals from entering a beacon receiver, provided that the interfering wavelengths are at least 10 per cent longer than the operating wavelength of the beacon. This difference is normally necessary because the filter must be matched to the regular transmission line over the beacon receiver band and the filter reactance changes rapidly with wavelength near the cutoff wavelength, $\lambda = 2a$.

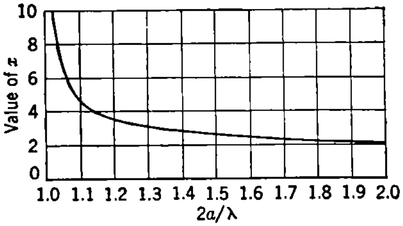


FIG. 7·18.—Attenuation in rectangular guide. Function $x$ plotted against $2a/\lambda$. $a$ = guide width. $b$ = guide height. $\lambda$ = wavelength (free space). Values of $x$ are for copper guide, $TE_{01}$ mode, and for a ratio $a/b = 2.24$. The attenuation is

$$\alpha = \frac{0.11x}{a^{3/2}}\ \text{db/foot.}$$

*Band-rejection Filters.*—If it is desired to keep interfering signals of a single frequency, or a narrow band, from entering the receiver of a beacon, a band-rejection filter is needed. A fairly closely coupled resonant cavity inserted so that its iris is in the broad face of a rectangular guide will offer a high impedance at resonance. The loaded $Q$ should be made as low as possible to minimize transmission losses subject to the requirement that the filter reactance be small at the beacon receiver band.

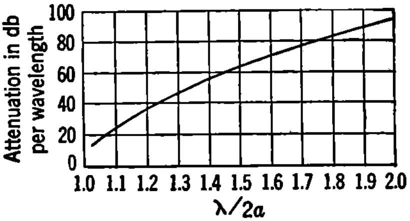


FIG. 7·19.—Attenuation in decibels per wavelength in rectangular guide beyond cutoff. $\lambda$ = wavelength. $a$ = width of guide. $TE_{01}$ mode.

Another type of band-rejection filter has been designed to attenuate 3-cm radiation in a 10-cm dielectric-filled coaxial transmission line. This is a section of line into which "chokes"—which are essentially resonant elements tuned to the unwanted frequency—are inserted. These are so spaced as to present either a short circuit (in parallel) or an open circuit (in series) for the undesired frequency, and to have little or no effect at the desired frequency. Relative attenuations of 10 to 15 db are readily attained.

**7·12. Radio-frequency Accessories.**[1]—In this section a few of the available r-f devices which have been of particular use in beacon design will be discussed briefly.

*Directional Couplers.*—A directional coupler may be used in a microwave beacon to extract a small amount of power from the transmission line for the purpose of measuring either the transmitted power or the frequency. It may also be used to inject a small amount of power into the receiving line for the purpose of monitoring the receiver. A directional coupler performs these functions without introducing an appreciable loss into the receiving or transmitting lines. It is superior to a fixed probe extending into the line because it gives a constant coupling independent of the standing-wave ratio in the line. This is true because it rejects the reflected wave—hence its alternative name, "wave selector." Each directional coupler has a fixed coupling determined by its design, and a suitable value for this coupling must be chosen for each application.

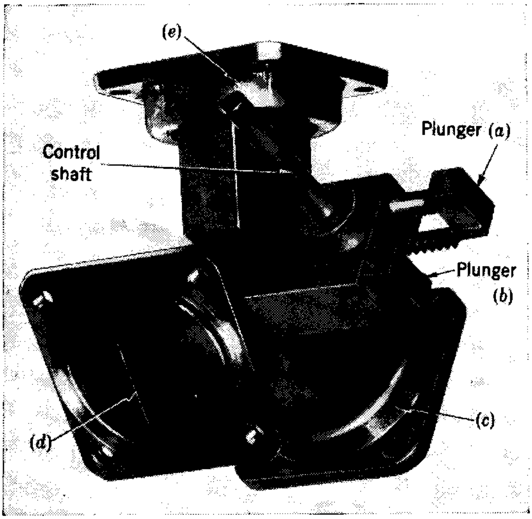


FIG. 7·20.—Three-cm r-f switch. With plungers in positions shown, (c) and (e) are connected and (d) is disconnected. With plunger (a) in and plunger (b) out, (c) and (d) are connected and (e) is disconnected.

In beacon receivers, directional couplers are used to couple the signal from the test set into the incoming r-f line in order to monitor the sensitivity, bandwidth, and pulse-width discrimination. One advantage of the directional coupler over the r-f probe in such uses is that the signal from the test set is directed to the receiver and only a negligible portion is allowed to go toward the antenna.

---

[1] By J. J. G. McCue, E. R. Gaerttner, and W. A. Downes.

*Dummy Loads.*—In high-power beacons, dummy loads are used both as a replacement for the antenna during testing and as a component part in the transmitter-stabilizing tuner. Dummy loads at 10 cm are available both in $\frac{7}{8}$-in. coaxial line and in waveguide. For 3-cm beacons, dummy loads are made of waveguide with a stepped polyiron insert. A dummy load must present a good match over a suitable frequency band and must be capable of dissipating the required amount of power.

At low powers (less than 1 watt average), a dummy load may consist of a dissipative platinized glass attenuator in a line terminated by a short circuit.

*Radio-frequency Switches.*—Radio-frequency switches are used when it is necessary to be able to switch r-f power quickly from one transmission line to another. In 3-cm beacons, they are convenient for connecting stand-by beacons to the antennas. A 3-cm switch is shown in Fig. 7·20.

At 10 cm, a multiposition r-f switch for coaxial line has proved useful in some applications. It was used in an airborne beacon, for example, to switch the antenna from one of two receiver inputs to the other; in this application it saved the use of an additional antenna on the aircraft.

*Phase Shifters.*—Phase shifters are used in beacons whenever it is necessary to adjust the effective r-f line length. For this reason they are often called "line stretchers."

In waveguide a phase shifter that utilizes a $\lambda/2$ section of dielectric that can be moved in the *H*-plane has been used. This unit does not change the physical length of the line, but it changes the wavelength by effectively varying the dielectric constant.

In $\frac{7}{8}$-in. coaxial line a "trombone" phase shifter which physically lengthens the line is available.

# CHAPTER 8

# BEACON RECEIVERS

BY E. R. GAERTTNER, J. H. TINLOT, AND G. P. WACHTELL

**8·1. Introduction.**[1]—The function of a beacon receiver is to detect the pulse-modulated r-f carrier received from an interrogator and amplify the demodulated signal to a useful output level. Beacon receivers differ from radar receivers in several respects. Usually they need not be as sensitive as radar receivers. They must often cover a wide frequency band. Their output need not necessarily provide as good pulse reproduction but simply a response suitable for tripping a trigger circuit, without excessive delay or variation of delay.

Because receiver characteristics change markedly over the range from 100 to 30,000 Mc/sec, it is necessary to divide the carrier frequencies into two regions. The r-f characteristics of ordinary grid-controlled tubes are such that they are no longer useful as r-f amplifiers above about 1500 Mc/sec.[2] The figure depends on the particular tube in question; furthermore, the criteria of usefulness are somewhat vague. Our treatment will be divided in accordance with the frequency. The classification is based on the use or absence of r-f amplifiers.

## GENERAL CONSIDERATIONS

BY E. R. GAERTTNER

**8·2. Types of Beacon Receiver.**—The types of receiver which are important in beacon applications are the superheterodyne (see Fig. 8·1), crystal video, tuned radio frequency, and superregenerative. In uhf superheterodynes, r-f amplifiers may be used. In the microwave region, none are as yet entirely suitable. The crystal-video receiver consists of a detector and a video amplifier, the last two components in Fig. 8·1. The trf receiver combines an r-f amplifier with a crystal-video receiver—components 1, 4, and 5 in Fig. 8·1. The superregenerative receiver

---

[1] By E. R. Gaerttner.

[2] It is, of course, true that any tube that oscillates at a frequency $f$ must amplify this frequency; otherwise it could not supply its losses and maintain oscillation. Radio-frequency amplification by klystrons, magnetrons, and other tubes at microwave frequencies is possible and has been accomplished in the laboratory, though not applied in the field. The signal-to-noise ratio is sometimes poor even when the gain is appreciable.

combines in one tube a special form of r-f amplifier and detector, and it usually has a video amplifier. The last two types of receiver cannot be used when r-f amplification is impractical.[1]

The fundamental properties which must be considered in the design of any beacon receiver are sensitivity, selectivity or frequency response, and fidelity.

**8·3. Sensitivity.**—Beacon receiver sensitivity is conveniently defined as the least r-f power which will produce a signal sufficient to trip a trigger circuit an arbitrary fraction of the time, usually taken to be 90 per cent. Since the maximum attainable sensitivity is needed for many applications, it is often necessary to work with signals near the noise level. All beacon circuits following the receiver which have so far been developed require a signal-to-noise ratio greater than 1. For example, it has been found from experience that for tripping a 2-μsec pulse-width discriminator reliably, a signal 8 db above the root-mean-square value of the noise is required.[2] There will, of course, still be some random tripping on noise at this level. The amount depends on the statistical distribution of the noise peaks and on the frequency spectrum of the noise. The effect of the latter depends upon the frequency response of the circuit that is triggered by the receiver output. Because of these factors, it is obvious that a numerical value for random noise tripping at the 8-db level, or at any other level, will not apply to all cases. If minimal noise tripping is essential, then an additional loss in sensitivity must be accepted.

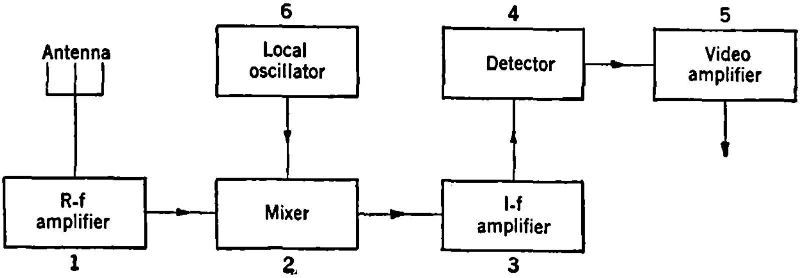


FIG. 8·1.—The essential components of a superheterodyne receiver. At microwave frequencies, the r-f amplifier is omitted. In a crystal-video receiver, components 1, 2, 3, and 6 are omitted; in a TRF receiver, components 2, 3, and 6 are omitted.

The sensitivity required for a given application depends on the system parameters and signal attenuation. A detailed account of these factors

---

[1] Microwave superregenerative receivers can be reasonably satisfactory; microwave trf receivers are not.

[2] Signals 8 db above noise are commonly called "tangential" because, on an A-scope, the noise superposed on the signal is roughly tangential to noise. Unsatisfactory as this criterion may appear, it is justified by custom and permits surprisingly accurate reproduction.

is given in Part I. In most cases, the power that will be available to trip the beacon at maximum range will lie between $10^{-7}$ and $10^{-12}$ watt.

The type of receiver to be used for a particular application depends on many factors. The choice is dictated first of all by the sensitivity requirements for a given range performance.

The maximum sensitivities in microwave beacon receivers (with bandwidths of 1 Mc/sec or more), can be obtained with the superheterodyne. They vary from $10^{-9}$ to better than $10^{-12}$ watt.

The crystal-video receiver, depending as it does on a square-law detector, is inherently much less sensitive to weak signals than a superheterodyne. Crystal-video receivers have sensitivities near $10^{-8}$ watt. It is evident that a good superheterodyne receiver can give many times the range performance of a good crystal-video receiver.

In a beacon system using a crystal-video beacon receiver, the range is not independent of the average power as it is to a first approximation when superheterodynes are used. The integrated pulse power detectable by a superheterodyne is roughly constant; if the peak power is doubled and the duration halved, doubling the bandwidth of the superheterodyne receiver will again give nearly the same signal-to-noise ratio. With a crystal-video receiver this is not true, since the output power will be doubled by making the same changes. Thus, in systems using crystal-video receivers, range performance can be increased without increasing the average transmitted power, by increasing the peak power and shortening the pulse in proportion.

Higher sensitivities can be obtained at lower frequencies; tuned-radio-frequency or superheterodyne receivers can have a sensitivity of $10^{-13}$ watt or better. The sensitivity of superregenerative beacon receivers runs from about $10^{-7}$ to $10^{-11}$ watt.

To achieve maximum sensitivity the gain of a beacon receiver must be sufficiently high to produce about 5 to 10 volts of noise at the output of the amplifier, since the following circuits in the beacon are found to work best for signals of about this amplitude.

**8·4. Frequency Coverage.**—Frequency coverage in a beacon receiver is defined as the range of frequencies over which the receiver gives a triggering signal for a signal power greater than a specified minimum. In other words, it is the frequency range over which the beacon receiver satisfies the sensitivity requirements discussed in the preceding section.

A narrow-coverage (high-selectivity) beacon receiver is one that responds to a single carrier-frequency channel or a narrow band of frequencies. It has a bandwidth comparable to the reciprocal of the pulse width (width of the region of the spectrum in which most of the energy of the pulses is found). Fortunately, it is not often necessary to strive for the maximum attainable sensitivity, as one must with a radar receiver.

Even in the so-called "narrow-coverage beacon receiver," it is generally desirable to increase the bandwidth somewhat, although some sensitivity may have to be sacrificed, to allow for variations of the interrogator frequency resulting from thermal changes, and so on.

A wide-coverage beacon receiver, on the other hand, is one which responds to a band of carrier frequencies which is very large compared with that needed for any one set of the pulse signals. Receivers that have very wide coverage are required because of the large scatter bands used by microwave radars. For several reasons, chief among which were the necessity of avoiding mutual interference between radars, the lack of tunable magnetrons, and the manufacturing tolerances of fixed-tuned magnetrons, 66 Mc/sec were allocated at 3300 Mc/sec and 110 Mc/sec at 9375 Mc/sec. A beacon must have a receiver broad enough to cover one of these bands, if it is to respond to all radars in it.

**8·5. Fidelity.**—The fidelity of a beacon receiver is a measure of its ability to reproduce the envelope of the pulse at the output of the video amplifier. To do this, the pass band of the amplifier must be comparable to the spread of the important frequency components of the pulse. For a video amplifier, this means that to reproduce a rise time of $\tau$ μsec, a pass band of $a/\tau$ Mc/sec is needed, where $a$ is about 0.35.

When nothing more than triggering of the beacon is required, faithful reproduction of the pulses is unnecessary. One important requirement, in these cases, is that the leading edge of the pulse be steep enough to trigger the following circuits with as little time delay as possible, in order to keep the range error small.

*Delay.*—The magnitude of the time delay depends on the frequency response of the amplifiers, both i-f and video, and on the signal strength. Intermediate-frequency amplifiers, which in beacon receivers usually have a wide pass band, cause very little delay. Video amplifiers, on the other hand, may cause appreciable time delay if the pass band is too small. The time delay varies inversely with the video bandwidth and its value for a given bandwidth depends on the signal strength. Thus, for 2-μsec signals slightly above the tangential level required for reliable triggering, the delay is negligible for a video pass band of 2 Mc/sec and is equal to the pulse width (2 μsec) for a pass band of about 200 kc/sec. For increasing signal strength the delay for the narrow pass band decreases somewhat.

*Discrimination.*—When information is to be derived from the pulse shape, however, as in pulse-width discrimination, the pulse must be reproduced with high fidelity so that the discriminator following the receiver can distinguish between pulses of different widths. The problem here is how to design a receiver that will preserve the pulse width over a wide dynamic range of signal strength, that is, one that will not unduly stretch the pulse for r-f signal strengths as much as 80 or 90 db above the minimum detectable level.

Superheterodyne receivers generally preserve the pulse width better than crystal-video receivers do. First, the mixer in a superheterodyne is linear, and the crystal detector in a crystal-video receiver is square law; for a variation of $n$ db in r-f power, there is a variation of $2n$ db in signal power in stages following the detector in the crystal-video receiver, and $n$ db in the superheterodyne. Second, a large part of the gain in a superheterodyne is in the i-f amplifier, which has good overload characteristics—that is, its recovery time is short in comparison with the duration of the pulse. On the other hand, because of its greater sensitivity, better overload characteristics are required to get pulse-width discrimination at short ranges. For example, assume sensitivities of 100 and 80 dbw, for the superheterodyne and crystal-video receivers respectively, and assume that discrimination is required to within $\frac{1}{10}$ mile of the beacon. If the signal received by the beacon at this range is 20 dbw, discrimination over a dynamic range of 80 db is required for the superheterodyne, and over a 60-db range for the video receiver measured at the antenna terminals, or 120 db at the first amplifier grid. Discrimination is readily obtained over the required dynamic range with a superheterodyne; it can be obtained, but with difficulty, in a practical crystal-video receiver.

## NARROW-COVERAGE RECEIVERS

**8·6. Superregenerative Receivers.**[1]—The superregenerative receiver is useful in beacon applications only when coverage of a moderately narrow band is desired. Most of the applications have been at lower frequencies, so the practical examples are taken from this region. Possible uses and limitations in the microwave region will also be discussed.[2]

The components of a superregenerative receiver are shown in the block diagram, Fig. 8·2. They consist of a superregenerative amplifier and detector, a quench oscillator, a quench filter, a video amplifier and (optionally) automatic gain control, AGC. The amplifier and detector, in conjunction with the quench oscillator, amplify and detect the r-f signal. The quench filter is a low-pass filter that prevents the quench frequencies from being amplified. Automatic gain control stabilizes the r-f gain of the receiver.

The receiver has some characteristics that make it desirable whenever small size and weight and moderately high sensitivity are required. The outstanding characteristic of the receiver is the high r-f gain of 50 db or

---

[1] Secs. 8·6 to 8·9 by E. R. Gaerttner.

[2] For a complete discussion of superregenerative receivers see *Microwave Receivers*, Vol. 23, Chap. 217, Radiation Laboratory Series.

more which can be obtained in one stage. Consequently, a receiver may consist of as few as three tubes: a detector, a quench oscillator, and a video amplifier. Its sensitivity is greater than that of a crystal-video receiver.

The superregenerative receiver can be used as a combination receiver and transmitter. It can be so used because the receiving tube, to be effective as a superregenerative amplifier, necessarily oscillates at the carrier frequency; consequently, it can be made to transmit on nearly the same frequency.

The superregenerative receiver is limited in use, however, in that it is not well suited to applications in which accurate range information is desired. This is because the time delay between the leading edge of the r-f pulse and the output video pulse depends on the time of arrival of the r-f pulse relative to the phase of the quench, and also on the r-f signal strength.

Perhaps the greatest disadvantage of the receiver is the instability of r-f gain, which is caused by changes in tube characteristics, variations in r-f loading by the antenna, and fluctuation in supply voltage. Stability can be obtained by adding an automatic gain control. However, the components that must be added to achieve the necessary stability tend to nullify in part the receiver's main features—its simplicity and small number of components.

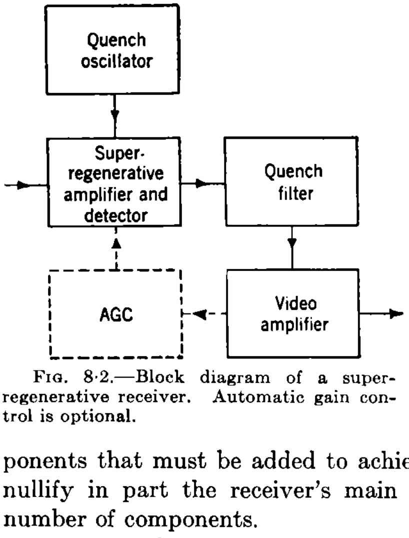


FIG. 8·2.—Block diagram of a superregenerative receiver. Automatic gain control is optional.

*Theory of Operation.*—It is well known that a tube connected so as to have sufficient regenerative feedback to produce self-sustained oscillation is also a sensitive receiver. The greatest sensitivity is obtained when the detector circuit is just able to sustain oscillations. Almost any oscillator, whether a triode, a velocity-modulated tube, or a magnetron,[1] can in principle be used as such a detector. In the superregenerative detector, the oscillations are not allowed to build up to a self-sustained state of constant amplitude, but are periodically quenched at a rate that is low compared with the normal oscillation frequency. The oscillations will then build up exponentially during the quench cycle; they are initiated

---

[1] An early Japanese radar set at 10 cm used a superregenerative magnetron receiver.

either by noise or by external signals. Two types of operation result, depending on when the oscillations are quenched. In the first, the quench action is applied before the oscillations build up to a saturation amplitude; in the second, the quench action is applied after the oscillations build up to a saturation amplitude. In the former, the rectified output is proportional to the r-f signal power; amplitude discrimination between signals and noise, therefore, is possible. In the latter, saturation amplitude is obtained independent of the signal strength, and the duration of the signal is proportional to the logarithm of the input power. After the output signal passes through the low-pass filter, its amplitude is proportional to the logarithm of the input signal.

*Sensitivity.*—Assuming sufficient r-f and video gain to bring the noise level to about 10 volts, the ultimate sensitivity depends upon the noise voltage generated at the input of the receiver. The noise factor is higher than that of a superheterodyne of comparable bandwidth. For example, the noise factor of an experimental 10-cm receiver using a 2C40 triode is about 30 db. The noise factor of a typical 150-Mc/sec superregenerative receiver with a bandwidth of 4 Mc/sec is 25 db. Thus, as far as sensitivity is concerned, the superregenerative receiver is often intermediate between the crystal-video receiver and the superheterodyne.

*Quench Frequency.*—The quench frequency is a small fraction of the carrier frequency. In practice it has been found that it should be about twice the modulation frequency. Thus, for a 2-$\mu$sec pulse, a minimum quench frequency of about 1 Mc/sec is required; for a 5-$\mu$sec pulse, 400 kc/sec, and so on. One reason for having a quench frequency higher than the modulation frequency, aside from the requirements of pulse reproduction, is that such a frequency can be filtered from the video amplifier by placing a low-pass filter between the detector and the amplifier.

Two factors determine the upper limit of the quench frequency. One is the point at which the gain is greatly reduced because the oscillations fail to build up sufficiently. This occurs for frequencies about equal to the r-f bandwidth. The other factor is the storage of energy in the antenna system, including the transmission line. If a long line is used, difficulty is encountered with any but perfectly matched lines, because power from the oscillation that was started in the previous quench cycle may be reflected back and forth and enough may remain when the quench cycle starts to interfere with proper operation. Antenna mismatches can likewise, by means of the "long-line" effect (see Sec. 13·9) prevent the receiver from operating at all at some frequencies.

In pulse detection, it follows from the frequency limitations that the period of the quench voltage should be less than the pulse duration. This inevitably leads to an unwanted time delay, however, which varies from zero up to the quench period. In addition, if the pulse duration extends over a number of quench periods, a series of signals that are separated in time by the quench period will be formed; the quench filter will tend to recombine them.

*Frequency Coverage.*—The r-f coverage of the superregenerative receiver is equal to the bandwidth of the superregenerative amplifier. It depends on a number of parameters, the most important of which are the transconductance ($g_m$) of the tube, the total shunt capacity of the resonant circuit, the strength of the input signal, and the waveform of the quench voltage. Sine-wave quench is almost universally used. A square-wave quench causes the oscillations to rise most rapidly and consequently gives the greatest bandwidth of any waveform (and also the poorest off-frequency rejection). For a $g_m$ of 5000 $\mu$mho and a shunt capacity of 20 $\mu\mu$f the approximate bandwidth at half amplitude for a square-wave quench is 5 Mc/sec. For other values of $g_m$ and capacity, the corresponding bandwidth can be scaled roughly to this example, according to the usual relation bandwidth = constant × $g_m/c$. A measured value for a beacon at 150 Mc/sec using a 2C22 tube is 4 Mc/sec.

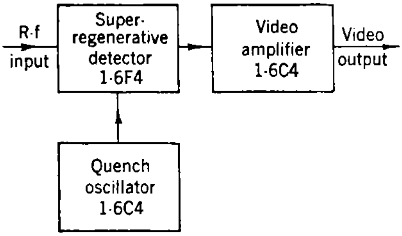


FIG. 8·3.—Uhf superregenerative receiver. It can be tuned from 660 to 720 Mc/sec. The quench frequency is 625 Mc/sec.

An example of a conventional superregenerative receiver with a sine-wave quench oscillator is shown in Fig. 8·3. The entire receiver consists of three tubes: a superregenerative detector, a quench oscillator, and one stage of video amplification. The sensitivity is 87 dbw and the frequency coverage for weak signals is 6 Mc/sec. The detector can be tuned to any frequency in the band 660 to 720 Mc/sec. The quench frequency is 625 kc/sec.

*"Single-cycle" Quenching.*—In applications in which it is intended that the superregenerative detector be insensitive except at a predetermined time, it is possible to use a "single-cycle" quench initiated by an external source. This type of operation has been used for coincident interrogation by pulses on two different carrier frequencies, as, for example, a microwave channel at 3000 Mc/sec and a uhf channel at 200 Mc/sec, the latter being detected by a superregenerator. The microwave pulse, after being received and amplified, initiates a square pulse that activates the superregenerative detector of the uhf receiver, thus rendering it receptive to a nearly simultaneous uhf pulse. The microwave pulse alone is ineffective because the superregenerative receiver receives no signal, and the uhf pulse alone is ineffective because the superregenerative detector is not sensitized.

In Fig. 8·4 is shown a block diagram of a coincident interrogation receiver using a single-cycle-quench uhf receiver. It consists of three tubes: a superregenerative detector, one stage of video amplification, and a quench generator. The quench is a square wave initiated by a 10-cm receiver in the manner described in the preceding paragraph. The sensitivity is 87 dbw and the frequency coverage is 2 Mc/sec.[1]

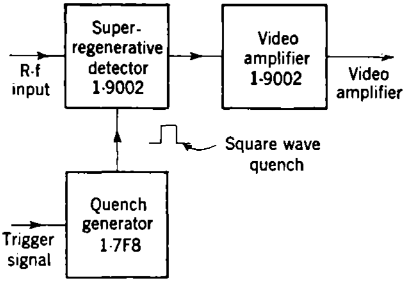


FIG. 8·4.—A single-cycle superregenerative receiver. It operates in coincidence with a 10-cm receiver which supplies a trigger to a square-wave quench-pulse generator for each 10-cm pulse received, thereby sensitizing the superregenerative detector for a short time. The sensitivity is 87 dbw and the band coverage is 2 Mc/sec. The detector can be tuned to any carrier frequency in the band 205 to 215 Mc/sec.

## 8·7. Superheterodyne Receivers.

—Superheterodyne receivers are characterized by high sensitivity and the possibilities of high selectivity. The high sensitivity is inherent in the type of detection used; high selectivity is possible because the frequency coverage, if one assumes a stable local oscillator, is determined at the relatively low intermediate frequency. Wideband coverage is likewise possible. In all other types of receivers, the frequency coverage is determined primarily by r-f components.

For convenience, the discussion of narrow-coverage superheterodyne receivers[2] is divided into two parts: (1) microwave receivers, which do not employ r-f amplifiers, and (2) lower-frequency receivers which, in general, do employ r-f amplifiers.

*Microwave Receivers.*—The sensitivity (for signal equal to noise) of an ordinary commercial superheterodyne radar receiver is about 120 to 130 dbw for an i-f bandwidth of 1 Mc/sec. Because of this high sensitivity, the superheterodyne is universally used in radar sets.

The frequency coverage required for operation on a single fixed channel is determined by the frequency stability of the interrogator transmitter and the beacon-receiver local oscillator, and by the bandwidth necessary to pass the video pulse satisfactorily. To achieve maximum stability AFC should be applied to both the interrogator transmitter and the beacon local oscillator.

*R-f Amplifiers.*—In the uhf region it is, in general, advantageous to add an r-f amplifier to the superheterodyne receiver. This is possible,

---

[1] For a detailed discussion of the circuits, see *Microwave Receivers*, Vol. 23, Chap. 21, Radiation Laboratory Series.

[2] For a complete discussion, see *Microwave Receivers*, Vol. 23, Radiation Laboratory Series.

at present, for frequencies less than about 1500 Mc/sec. There is, of course, no sharp line of demarcation since the use of an r-f amplifier depends upon getting tubes with suitable r-f characteristics.

A superheterodyne receiver, in conjunction with an r-f amplifier, affords a means of obtaining the ultimate sensitivity with good selectivity. The sensitivity depends on a number of factors, especially the noise factors of the mixer and i-f amplifier and the noise factor and gain of the r-f amplifier. It is, of course, clear that the noise factor of the r-f amplifier must be appreciably less than that of the mixer and i-f amplifier combined, if the ultimate sensitivity of the receiver is to be increased by adding an r-f amplifier. Moreover, considerable r-f gain is needed.

If the noise factor of the r-f amplifier is $F_1$, its gain (in power) $g_1$, and the noise factor of the mixer and i-f amplifier combined is $F_2$, the noise factor $F_{12}$ of the receiver using the r-f amplifier is[1]

$$F_{12} = F_1 + \frac{F_2 - 1}{g_1}. \tag{1}$$

As an example, suppose an r-f amplifier 2 Mc/sec wide has a noise factor of 3 (5 db) and a power gain of 20 (13 db). If the noise factor of the mixer and i-f amplifier is 10 (10 db), the noise factor of the receiver using the r-f amplifier will be $3 + \frac{9}{20} = 3.5$, or 5.4 db. The improvement is then 4.6 db.

The selectivity curve of the receiver is the product of the selectivity curves of the r-f and i-f amplifiers. Since i-f bandwidth can be made small, the problem of getting high selectivity is simple, as compared with the same problem in the trf receiver, for example.

It is clear from the sensitivity and selectivity characteristics that this type of receiver finds its best use in highly selective systems. Its complexity does not recommend it for use in systems in which low sensitivity and poor selectivity are acceptable.

**8·8. Crystal-video Receiver with R-f Filter.**—Although a crystal-video receiver is inherently a wide-coverage receiver, its frequency coverage can nevertheless be narrowed by inserting a bandpass filter in the r-f line between the antenna and the crystal detector. This system is simpler than the narrow-coverage superheterodyne and is equally—or, perhaps, even more—stable in frequency. A good filter should not introduce more than a 1- or 2-db loss in sensitivity.

The limitation of this system is the low sensitivity of the crystal-video receiver, which is currently 75 to 85 dbw in the microwave region. Nevertheless, it has a number of applications.

A typical example is one in which a high-power microwave ground radar is used for surveillance and control of aircraft equipped with bea-

---

[1] See, for example, *Vacuum Tube Amplifiers*, Vol. 18, Chaps. 13, 14, and 15, Radiation Laboratory Series.

cons. Since the pulse power and antenna gain of such a ground system are large, the beacon sensitivity may be low, even for long-range performance. It can readily be shown that the required sensitivity for reliable operation up to 200 miles with the usual 10-cm ground-control radar is approximately 60 dbw, a value readily obtainable with a crystal-video receiver.

**8·9. TRF Receiver.**—The combination of an r-f amplifier with a crystal-video receiver constitutes what is called a "tuned radio-frequency receiver." The use of this type of receiver is naturally restricted to frequencies at which r-f amplifiers are practical.[1]

The sensitivity of a trf receiver may be made comparable to that of a superheterodyne if r-f amplifiers with suitable noise factor and gain are available for the frequency range considered. For example, the noise power of an r-f amplifier at a few hundred megacycles per sec, of 5-Mc/sec bandwidth and 10-db noise factor, may be, say, 37 db less than the noise power (not the triggering level) of a good crystal-video receiver, which is about $10^{-9}$ watt. For the minimum detectable power to be substantially equal to the noise power of the r-f amplifier, the r-f gain should be sufficient for the amplified r-f noise to exceed the video noise by about 10 db. For this to be true, the r-f gain must be 47 db. For larger bandwidths and higher noise factors the required gain is, of course, less.

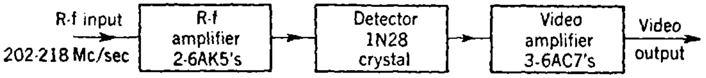


FIG. 8·5.—A 200-Mc/sec TRF receiver. It has a sensitivity of 130 dbw and an r-f bandwidth of 3 Mc/sec. The r-f amplifier has 30 db gain. The video amplifier is 0.6 Mc/sec wide and has 116 db gain.

The frequency coverage is limited by the r-f response of the amplifier. On the one hand, high selectivity cannot be achieved because narrow-band, high-gain r-f amplifiers are unstable. On the other hand, coverage comparable to that of a crystal-video receiver cannot be obtained. Consequently, for covering a band of frequencies this type of receiver is intermediate between the crystal-video and the superheterodyne.

The design of a tunable multistage r-f amplifier at high frequencies is relatively easy if each stage is tuned independently. It becomes difficult if the tuning must be accomplished with a single knob and the several stages must, therefore, track.

The application most suitable for this type of receiver is one in which high sensitivity and narrow coverage (but not high selectivity of the sort obtainable with a superheterodyne) are required. Economy in tubes and power is possible with most of the receiver gain in the video amplifier.

---

[1] For a further discussion see *Microwave Receivers*, Vol. 23, Chap. 5, Radiation Laboratory Series.

The receiver does not have the difficulty with image response experienced with the superheterodyne.

In Fig. 8·5 is shown a block diagram of the trf receiver used in a shipborne beacon. The entire receiver consists of two stages of r-f, a crystal detector, and three stages of video amplification. The noise power is 130 dbw. The r-f amplifier has a bandwidth of 3 Mc/sec, 30 db r-f gain, and it can be slug-tuned to a particular carrier frequency within the range 202 to 208 Mc/sec.[1]

**8·10. Radio-frequency Channeling.**[2]—We have seen in Sec. 5·14, that one of the simplest kinds of interrogation coding that can be applied to a beacon system is the use of several different radio frequencies for interrogation. This implies that the beacon will be equipped with a narrow-coverage receiver capable of sufficient selectivity to receive any one of these channels and to reject the others. The receiver must also be capable of being switched readily from one channel to another.

The various receivers that have been discussed are all capable of fulfilling such requirements; they must, nonetheless, be considered carefully during the design of the receiver. The over-all system design will dictate the amount of adjacent-channel selectivity required and the necessary ease of switching.

*Adjacent-channel Selectivity.*—This may be defined as the quantity $P_a/P_b$, expressed in decibels, where $P_a$ is the amount of received power, of frequency corresponding to channel $a$, which is required to trigger a beacon tuned to the adjacent channel $b$, and which is capable of being triggered by an amount of received power $P_b$ in channel $b$.

In a superheterodyne receiver, this quantity can be made very large indeed; 40 db or more is readily attained. One must take care that the intermediate frequency—or channel separation—is so chosen that the adjacent channel does not fall near the image frequency of a superheterodyne without good image rejection. In superheterodynes using r-f amplifiers this is easier to accomplish.

In trf receivers, the adjacent-channel selectivity will depend entirely upon the selectivity and number of the r-f amplifier stages. High selectivity is likely to involve difficulty in switching from one channel to another.

In a crystal-video receiver with an r-f filter, the r-f selectivity is determined entirely by the filter; a filter with suitable properties can be chosen.

The r-f selectivity of the standard sine-wave quenched superregenerative receiver is good; it corresponds to that of an i-f amplifier with an infinite number of stages. (See *Microwave Receivers*, Vol. 23, Chap. 21.)

---

[1] For a detailed discussion of the circuits see *Microwave Receivers*, Vol. 23, Chap. 5, Radiation Laboratory Series.

[2] By E. R. Gaerttner and G. P. Wachtell.

*Channel Switching.*—The process of tuning the receiver from one frequency to another varies in complexity with the type of receiver being used. In the crystal-video receiver with an r-f filter and in the superregenerator, only one component needs to be tuned. In a microwave superheterodyne, only one receiver component, the local oscillator, needs to be tuned if the other r-f components have sufficient bandwidth. In a uhf superheterodyne, the r-f amplifier and mixer as well may need to be tuned. In the trf receiver the r-f amplifier needs to be tuned. The tuning in the last two cases is apt to be time-consuming.

## WIDE-COVERAGE RECEIVERS: SUPERHETERODYNES[1]

The following discussion of wide-coverage receivers is restricted to microwave receivers, because the problem of wide frequency coverage has so far arisen only in the microwave region.

In the microwave region, r-f amplifiers are not available. Furthermore, the mixer is generally a crystal instead of a tube because the noise factors of crystals are considerably lower. There are exceptions; some triodes, for instance the 2C40 lighthouse tube, are moderately satisfactory up to 3000 Mc/sec. A mixer made with this tube is discussed in Sec. 8·15.

A superheterodyne receiver is used in applications requiring greater receiver sensitivity than can be obtained with the simpler crystal-video receiver. The use of superheterodynes is necessary whenever maximum range performance is a requirement.

**8·11. General Considerations.** *Echo Suppression.*[2]—In pulse-width discrimination it is necessary for the receiver to preserve the pulse shape well enough for the discriminator to distinguish between 1-μsec search pulses and 2-μsec beacon-interrogation pulses. This requirement, of course, imposes certain restrictions on the fidelity of the receiver, especially the pulse stretching due to poor overload characteristics of the video amplifier.

In addition, however, there is the possibility that the effective pulse length may be increased by r-f echoes from near-by objects. Thus, a 1-μsec pulse and its echo from a near-by hill or building may be superimposed; if the echo lags the direct pulse by 1 μsec, an effective 2-μsec signal is received. The direct signal is usually stronger than the echo; however, the combination may still be amplified sufficiently to activate the discriminator. Consequently, some provision for suppressing the echoes must be made in the amplifier if good pulse-width discrimination is desired, especially in a sensitive receiver like the superheterodyne.

---

[1] By E. R. Gaerttner.

[2] By J. H. Tinlot and E. R. Gaerttner.

Because the echoes necessarily occur after the main signal and are of lesser amplitude, the problem resolves itself into desensitizing the amplifier for the period of time in which undesired echoes are expected.[1]

*Methods of Covering Microwave Bands.*—Frequency coverage is usually limited by the bandwidth of the i-f stages because the bandwidth of the mixer and r-f components is, in general, much larger. With conventional methods of coupling i-f stages, the bandwidths are relatively narrow—usually less than 8 or 10 Mc/sec; consequently, to cover the microwave beacon bands, a number of special methods have had to be used. Three methods that have been successfully applied to superheterodynes are the following.

1. Single-frequency local oscillator with wideband i-f amplifier.
2. Square-wave-modulated (switched-frequency) local oscillator.
3. Frequency modulation of local oscillator.

To keep the problem concrete, it should be remembered that the frequency coverages required of 10-cm and 3-cm beacon receivers are 66 and 110 Mc/sec, respectively. The methods for achieving wide frequency coverage are all designed to provide coverage of 70 Mc/sec or more.

**8·12. Wideband I-f Receiver.**[2]—The wideband i-f amplifier is a natural extension of the narrow-band i-f amplifier to give wide coverage in one sideband. To accomplish this, the conventional i-f interstage coupling methods are replaced by another type known as "shunt-series peaking." Essentially, the shunt-series peaking circuit is a video circuit that may be designed to give a pass band extending from zero cps to a maximum frequency determined by the filter constants and the shunt capacities of the tubes. In order to prevent the i-f pass band from overlapping the pass band of the video amplifier, suitable low-frequency rejection filters are introduced. These filters may be in the cathodes of the i-f tubes (resulting in selective degeneration) or they may be added to the interstage coupling filter, or both.

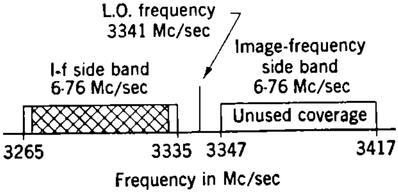


FIG. 8·6.—Coverage of the 10-cm band from 3267 to 3333 Mc/sec with a wideband i-f receiver. The required coverage is cross-hatched. Coverage is accomplished by making the pass band of the i-f amplifier greater than the width of the scatter band and by adjusting the local-oscillator frequency so that all portions of the band are covered. The image-frequency sideband provides undesired coverage.

---

[1] For a detailed discussion of the "echo-suppression" circuit as applied to i-f amplifiers see *Microwave Receivers*, Vol. 23, Chap. 19, Radiation Laboratory Series. An application of this type of circuit is given below in Sec. 8·13.

[2] Secs. 8·12 to 8·15 by E. R. Gaerttner.

In a receiver designed to cover the 66-Mc/sec band at 10 cm, the i-f amplifier must have a pass band of at least 66 Mc/sec. The i-f amplifier shown in the following example has a pass band extending from 6 to 76 Mc/sec. The method of covering the band with this amplifier is illustrated in Fig. 8·6. The image-frequency band, also shown in Fig. 8·6, lies outside the desired frequency coverage and is not used. It may even be a source of interfering signals.

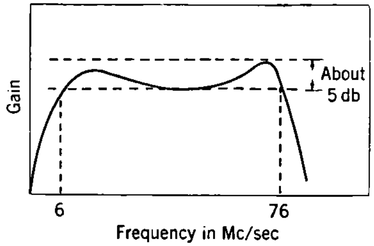


FIG. 8·7.—Frequency response of a 6- to 76-Mc/sec i-f amplifier using shunt-series peaking.

In principle it is desirable for the frequency response of the i-f amplifier to be perfectly flat; actually the curve is like that shown in Fig. 8·7. The variation of 4 to 6 db in gain over the band is not objectionable if the sensitivity is adequate at all frequencies. At the lowest point, the triggering sensitivity is greater than 100 dbw.

High fidelity, necessary for pulse-width discrimination, is obtained by using a video amplifier with a bandwidth of about 2 Mc/sec. This is possible in this type of receiver because of the relatively high frequency of cutoff at the lower limit of the i-f pass band (6 Mc/sec in the example). Thus ample bandwidth is allowed in the video without overlapping into the i-f pass band.

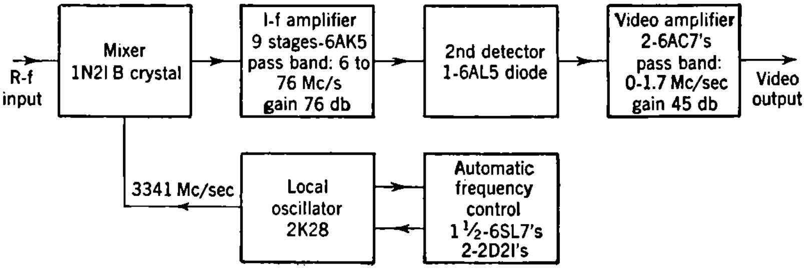


FIG. 8·8.—The wideband i-f receiver used in AN/CPN-17. This receiver covers the band 3265 to 3335 Mc/sec with a minimum sensitivity of 100 dbw. The band coverage is stabilized by AFC of the local oscillator.

The ultimate limit of the band coverage that can be achieved by this method is determined by the greatest width of the i-f pass band that can be achieved. With the best tubes (6AK5) and circuits now available this is about 100 Mc/sec.

Figure 8·8 shows a block diagram of such a receiver used in the AN/CPN-17, a high-power 10-cm beacon for ground and shipborne installations. The specifications call for a minimum sensitivity of about 95 dbw and pulse-width discrimination. No severe restrictions are placed on size, weight, or power consumption. Some features of this receiver are AFC of the local oscillator and duplexing; AFC is discussed in Sec. 8·16, and duplexing in Sec. 7·9.

**8·13. Square-wave-modulated Local Oscillator.**—This method is used when the band to be covered is larger than the maximum practical i-f bandwidth, and both high sensitivity and high fidelity are necessary. The wideband i-f receiver is thus ruled out. The method of coverage is given in Fig. 8·9.

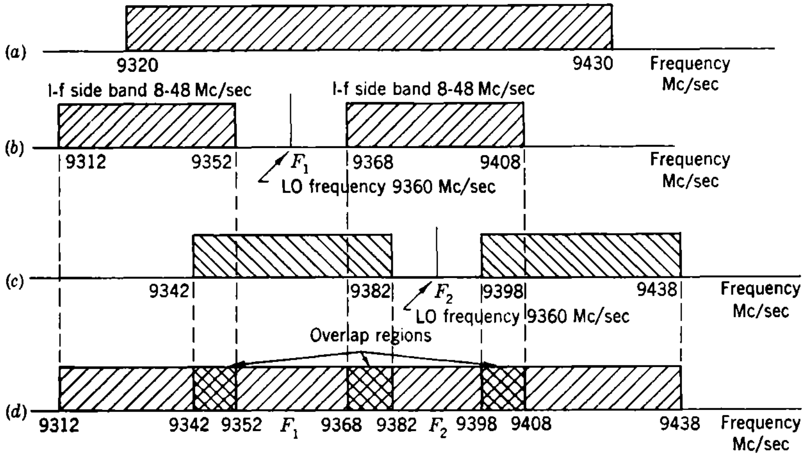


Fig. 8·9.—The method of covering the 9320- to 9430-Mc/sec band by means of a superheterodyne receiver with an i-f pass band of 8 to 48 Mc/sec and a square-wave-modulated local oscillator. (*a*) The required coverage, 9320 to 9430 Mc/sec. (*b*) The pattern of the band covered with the LO frequency 9360 Mc/sec. (*c*) The pattern of the band covered with the LO frequency 9390 Mc/sec. (*d*) The sum of (*b*) and (*c*)—the total coverage. All parts of the band, except for regions of overlap (cross-hatched) are covered half the time. The switch frequency for the local oscillator is 200 cps.

The frequency coverage of the receiver is 126 Mc/sec, although the actual bandwidth of the i-f amplifier is only 40 Mc/sec. The i-f stages use shunt-series coupling with a pass band extending from 8 to 48 Mc/sec. By using the image-frequency bands, therefore, signals can be received with frequencies either 8 to 48 Mc/sec above or 8 to 48 Mc/sec below the local-oscillator frequency. Thus, a band of frequencies 96 Mc/sec wide, with a 16-Mc/sec gap in the middle, is covered. To fill in the gap and widen the over-all band coverage, the frequency of the local-oscillator tube is switched one-half the time to a frequency 30 Mc/sec away. The four bands thus cover a total of 126 Mc/sec. The two bands covered by having the local oscillator at one frequency overlap the two bands covered with the local oscillator at the other frequency. The overlap regions, as shown in the illustration, are 10, 14, and 10 Mc/sec wide, respectively. If the frequency of a received signal lies in one of these overlap regions, the signal will get through the receiver all the time. If the signal frequency is within the band but does not lie in one of the overlaps, the signal will get through the receiver only half the time.

This alternation of intervals of sensitivity and insensitivity is the major disadvantage of the square-wave-modulated local oscillator as a means of achieving wide coverage. By using the image-frequency bands, a maximum of four times the i-f coverage can be achieved; but the receiver will then be sensitive to any one frequency no more than half the time.

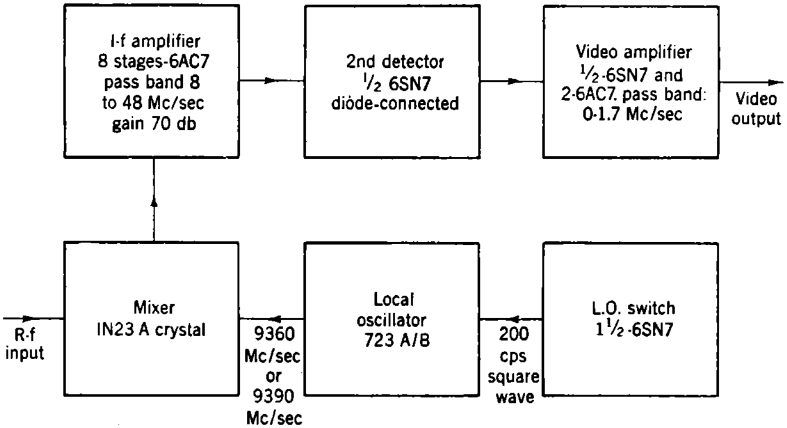


FIG. 8·10.—Superheterodyne receiver using image-frequency response and a square-wave-modulated local oscillator to achieve wide coverage. The coverage is obtained in the manner shown in Fig. 8·9.

The 50 per cent response introduces a scanning loss of 1.5 db in the response link, as pointed out in Sec. 2·6. In order to lessen the effect of the switching, the i-f pass band is usually extended toward the local-oscillator frequency to increase the region of overlap, as in the example given. The switching rate is made somewhat less than one-half the minimum repetition rate of the interrogating signals, or about 150 to 200 cps. High fidelity, necessary for pulse-width discrimination, is obtained by using a video amplifier with a bandwidth of about 2 Mc/sec.

Figure 8·10 shows a block diagram of such a receiver used in the AN/CPN-6, a high-power 3-cm beacon designed for ground and shipboard use. The specifications call for a minimum sensitivity of 90 dbw and pulse-width discrimination. The actual sensitivity runs between 100 and 110 dbw. No severe limitations are placed on size, weight, or power consumption. One feature of the i-f amplifier is echo suppression, discussed above in Sec. 8·11.

**8·14. Sine-wave Frequency-modulated Local Oscillator.**—This type of receiver is used in applications in which greater sensitivity than can be attained with a crystal-video receiver is required, but pulse-width discrimination is not required. It provides the widest coverage of all superheterodyne methods.

In the wideband i-f receiver discussed above, the frequency coverage is determined by the i-f pass band. It is clear that if the i-f amplifier could be made to amplify frequencies from zero to $f$, then the total band coverage would be $2f$ because the two adjacent sidebands could be employed as shown in Fig. 8·11.

In practice, of course, it is not possible to extend the i-f pass band to zero frequency. Thus a "hole" in the response curve will appear at the local-oscillator frequency, the width of the hole being twice the low-frequency cutoff of the i-f pass band. If the frequency width of the hole is made less than the frequency-spectrum-width of the received pulse, a pulse that is centered on the hole is still amplified although with reduced gain and considerable distortion.

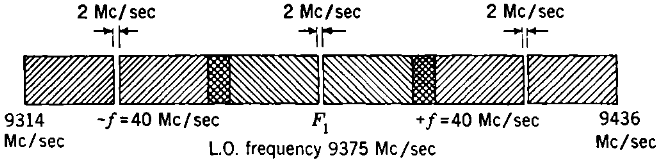


Fig. 8·11.—Coverage of the 9320- to 9430-Mc/sec band by a superheterodyne receiver using a frequency-modulated local oscillator. Local-oscillator frequencies of 9335, 9375, and 9415 Mc/sec are produced by modulating a local oscillator on 9375 Mc/sec with a 40-Mc/sec sine wave. The i-f pass band extends from 1 to 21 Mc/sec. The 2-Mc/sec "holes" are filled by "wobbling" the local-oscillator frequencies over 4 Mc/sec with a small amount of power-line frequency f-m.

The coverage of such a receiver is, of course, limited by the maximum frequency response of the i-f amplifier. If greater frequency coverage than can be attained in this way is desired, it can be had by providing additional local oscillators. To understand the principle of such a receiver, assume that three local oscillators, separated by 40 Mc/sec, furnish power to the same crystal mixer. If the i-f pass band extends from 1 Mc/sec to 21 Mc/sec, the r-f coverage would then be as indicated in Fig. 8·11. It is clear that 122 Mc/sec is covered all the time (except for the three 2-Mc/sec gaps) by a receiver using an i-f amplifier of about one-sixth this total bandwidth, with some overlapping of bands. To fill in the 2-Mc/sec gaps, a small amount of audio-frequency FM (total deviation 4 Mc/sec) is applied to the local oscillators. This is obtained by applying a small a-c voltage, derived from the power line, to the reflector. The beats between the local oscillators lie at multiples of 40 Mc/sec, outside the i-f pass band.

In the sine-wave frequency-modulated receiver the three local-oscillator frequencies are obtained from a single tube by frequency-modulating a velocity-modulated local oscillator. This is accomplished by the application of a sine-wave voltage to the reflector. It is well known that, if an oscillator of frequency $F$ is frequency-modulated by a sine-wave of frequency $f_m$, discrete frequencies $F \pm nf_m$ among others will appear simultaneously in the output ($n$ is a positive integer). Consequently, to provide for the local oscillators shown in Fig. 8·11, 40-Mc/sec frequency modulation is required. The number and relative amplitudes of frequencies appearing depend upon the frequency deviation used. In practice these parameters depend upon the characteristics of the local oscillator tube.

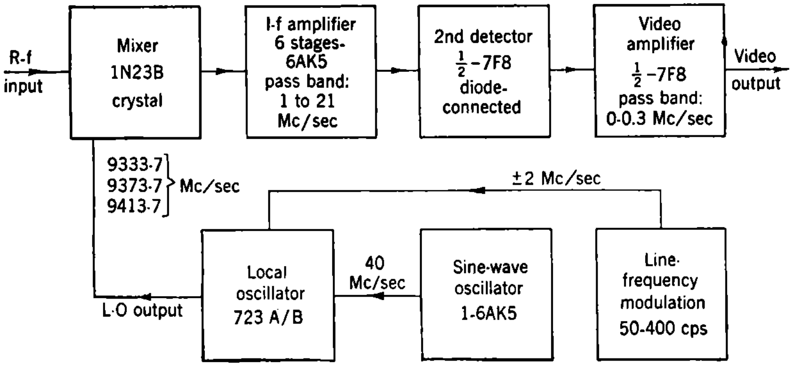


FIG. 8·12.—Block diagram of a superheterodyne receiver using a sine-wave frequency-modulated local oscillator as shown in Fig. 8·11.

The video pass band is made very narrow (less than 0.3 Mc in the example, Fig. 8·12) to avoid overlapping into the i-f pass band; consequently 2-μsec beacon interrogation pulses are distorted so badly that pulse-width discrimination is impossible. However, a signal suitable for triggering the succeeding beacon circuits is produced.

Figure 8·17 shows a block diagram of a complete receiver using local-oscillator frequency modulation. Figure 8·13 is a photograph of such a receiver used in AN/UPN-3, a portable 3-cm beacon. The specifications call for a minimum sensitivity of 90 dbw and no pulse-width discrimination. Emphasis is placed on light weight, low power consumption, and ease of adjustment. A complete circuit diagram of the receiver and an analysis are given in *Microwave Receivers*, Vol. 23, Chap. 19.

This type of receiver provides the widest coverage for a given i-f bandwidth. In the example given, and in actual practice, only three local-oscillator frequencies are used. Actually, because of the characteristics of f-m sidebands, other virtual local oscillators are present at ±80 Mc/sec from the center frequency, others at ±120 Mc/sec, and so on, with less and less power. In the receiver described, enough local-oscillator power is present so that two additional 40-Mc/sec-wide regions exist, one on either side of the main frequency-coverage band; receiver sensitivity in these is down only 10 to 15 db as compared with the central region.

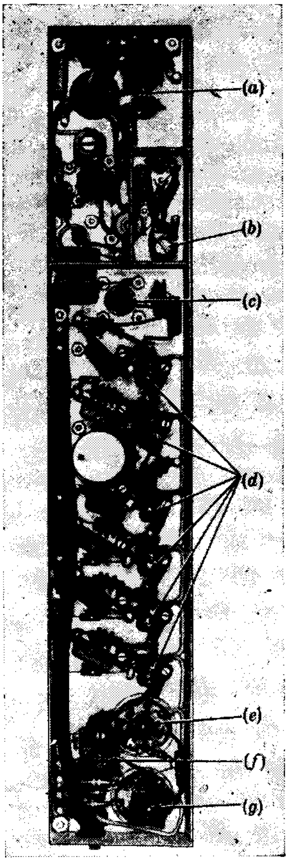


FIG. 8·13.—Lightweight, low-power receiver strip using frequency-modulated local oscillator, designed for the portable 3-cm beacon, AN/UPN-3. (a) Leads to power terminals. (b) 40-Mc/sec oscillator. (c) I-f input. (d) I-f amplifiers. (e) Second detector. (f) Video amplifier. (g) Multivibrator (output trigger).

Increasing the coverage by actually providing more local oscillators is no doubt feasible. This would reduce the sensitivity somewhat by increasing the crystal current in the mixer above the optimum, but the reduction would not be great.

It is possible to create the sideband power needed for virtual local oscillators by applying the modulating voltage to the crystal mixer rather than to the local oscillator. By such methods, higher modulating frequencies can be used than can successfully be applied to local-oscillator tubes.

If the i-f amplifier has a pass band from 1 to 100 Mc/sec, one local oscillator frequency-modulated at 200 Mc/sec (if this were feasible) would give a coverage of 600 Mc/sec by the methods already used. If three such modulated LO's were used, a coverage of 1800 Mc/sec could be obtained. These coverages are sufficiently great to require reexamination of the selectivity of r-f components.

**8·15. Mixers.** *Crystal Mixers.*—The mixer[1] in a superheterodyne receiver is the component in which the r-f signal is mixed with the local-oscillator frequency to produce the intermediate frequency.

---

[1] For a complete discussion of mixers see *Microwave Receivers*, Vol. 23, Chap. 2, Radiation Laboratory Series.

At 10,000 Mc/sec and higher frequencies, it may consist of a section of waveguide suitable for the frequency range considered, with a crystal mounted across the guide. At lower frequencies the mixer may consist of a section of coaxial line. In either case, the line must be properly terminated to prevent r-f loss. This requires a tuning adjustment if maximum power sensitivity is desired for all crystals of a given type. With a fixed-tuned adjustment, up to 1 db in sensitivity may be lost with some crystals. The mixer must also be sufficiently broadband to cover the frequency band used. This requirement is, in general, easy to meet.

*Mixer Crystals.*—Two quantities which are measures of the performance of a mixer crystal[1] are the conversion loss and output noise ratio (noise temperature). The output noise ratio indicates how much more noise is generated by the crystal than by an equivalent resistance at the same temperature. Numerical values for a number of crystals in common use are given in Table 8·1.

TABLE 8·1.—CHARACTERISTICS OF MIXER CRYSTALS

| Crystal | Band, cm | Max. conversion loss, db | Output noise ratio, max. |
|---|---|---|---|
| 1N21 | 10 | 8.5 | 4.0 |
| 1N21A | | 7.5 | 3.0 |
| 1N21B | | 6.0 | 2.0 |
| 1N23 | 3 | 10.0 | 3.0 |
| 1N23A | | 8.0 | 2.0 |
| 1N23B | | 6.5 | 2.0 |

*Tube Mixers in the Microwave Region.*—Some triodes may be used as mixers in wide-coverage 10-cm superheterodynes. At first sight it appears that a tube mixer with its high noise factor (about 10 db more than a crystal) would give an unwarranted loss in sensitivity over what might be obtained with a crystal mixer. However, the noise factor of a wideband i-f amplifier may be high, perhaps as much as 10 db. Thus, with a crystal mixer, the noise level might be primarily that of the i-f amplifier. Consequently, there is little loss in sensitivity with a tube mixer as long as its noise factor is not much greater than that of the i-f amplifier. Moreover, triodes give conversion gain (10 db in one 10-cm experimental model), thus overbalancing i-f noise and requiring less gain in the i-f stages than would be required for a crystal mixer.

From Eq. (1), Sec. 8·7, we can see just what loss in performance may be expected in a representative case. Suppose we have a crystal mixer

---

[1] Mixer crystals are discussed in detail in *Crystal Rectifiers*, Vol. 15, Radiation Laboratory Series.

with a gain $g_1$ of $\frac{1}{4}$ (conversion loss 6 db) and an output noise ratio of 2.0. Its noise factor $F_1$ is then 4·2 or 8. Suppose the noise factor $F_2$ of the wideband i-f amplifier is 10 (10 db). The noise factor of the receiver is then $8 + \frac{10 - 1}{\frac{1}{4}} = 44 = 16.5$ db.

If we now substitute for the crystal mixer a tube mixer whose noise factor is 100 (20 db), but which has a conversion gain $g_1$, of 10, the noise factor of the receiver will be $100 + \frac{10 - 1}{10} = 101 = 20$ db. Thus the tube mixer reduces the performance by only 3.5 db in this case.

A 10-cm mixer cavity using the 2C40 lighthouse tube is described in Sec. 7·10. The noise factor of such a mixer is about 100 with 20 mw of local-oscillator excitation. Because of the tight coupling and electron loading the $Q$ is about 10, assuring adequate coverage of the 66-Mc/sec band.

One useful characteristic of the triode mixer is that antenna duplexing can be combined with mixing in the same component, as we have seen in Sec. 7·10.

**8·16. Receiver AFC.**[1]—Automatic frequency control[2] is desirable in a beacon superheterodyne receiver to assure that the receiver covers the specified band of frequencies at all times. In the superheterodyne receiver, which in the majority of cases covers the required frequency band without much excess coverage, the need for AFC arises from changes in frequency of the local oscillator resulting from thermal drifts, aging, line-voltage changes, and so on. The function of an AFC circuit is to prevent drift of the local oscillator. Although drift can be minimized by careful construction of the receiver, AFC eliminates not only the possibility of its passing unnoticed but also the necessity for manual control of the tuning—an important point in assuring ease and reliability of operation.

*The Phase Discriminator.*—Like all automatic control systems, AFC systems depend for their operation on the generation of an error signal. In a beacon receiver, it is required that the local oscillator be tuned to be as close as possible to a specified frequency. With microwaves, this standard frequency can be set most easily by means of a fixed resonant cavity (see Fig. 8·14). The error signal required for AFC may be obtained by modulating the frequency of either the local oscillator or the cavity. The error signal thus obtained should not only indicate deviation from the correct frequency but also the sign of the deviation. Since it is much simpler to produce the desired modulation of the local oscillator

---

[1] By J. H. Tinlot and E. R. Gaerttner.

[2] A complete description of receiver AFC is given in *Microwave Receivers*, Vol. 23, Chap. 30, Radiation Laboratory Series.

than it is to modulate the frequency of the standard cavity, only the former means will be discussed.

The phase discriminator circuit to be described provides a control signal that acts only when the oscillator is tuned to a frequency slightly lower than that of the cavity. If the oscillator is tuned higher than the cavity frequency, no control (or error) signal will result; other means of correcting such a deviation are provided.

A means of detecting on which side of the cavity frequency the oscillator is tuned is provided by sine-wave frequency modulation of the oscillator. Suppose the frequency of the oscillator to be momentarily slightly higher than that of the cavity, and suppose that a little frequency modulation is applied to the local oscillator. The signal appearing at the output of the cavity crystal will be a superposition of a rectified d-c voltage and an a-c component, because the oscillator is tuned to a frequency corresponding to a sloping portion of the cavity response curve. If the local oscillator is at a frequency slightly lower than that of the cavity, the same situation results, but with the important exception that the a-c component will now be exactly 180° out of phase. It is the difference in phase between the modulating voltage and the a-c component of the error signal that is used to determine whether the oscillator is tuned higher or lower than cavity frequency.

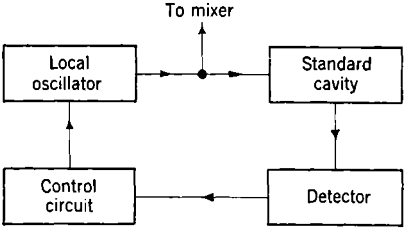


FIG. 8·14.—Block diagram of receiver AFC circuit.

*Gas-filled-tube Control Circuit.*—One widely used control circuit employs two 2D21 gas-filled tubes. If we neglect for the moment the action of the first tube, the second, or search, tube causes the local oscillator to change in frequency at a slow rate from higher to lower frequencies. A further slight frequency modulation is effected by a superimposed 60-cps voltage. A sawtooth oscillation is produced at the plate of the gas-filled tube with a frequency of about 3 cps. When it is applied to the local-oscillator reflector through an isolation circuit, the local-oscillator frequency also varies along a sawtooth path. Thus, the local-oscillator frequency searches for the cavity frequency.

The amplified output of the crystal detector of the reference cavity is applied to the control grid of the first gas-filled tube, and a 96-volt, 60-cps signal is applied between its screen grid and cathode. When the local oscillator is tuned far from the cavity frequency or is even slightly above the proper frequency, the tube remains nonconducting and does not affect the sawtooth oscillation of the search tube in any way. When the local oscillator is tuned slightly below the cavity frequency as a result of the search of the second tube, both the control grid and the screen grid are driven positive in phase and cause the tube to conduct. Small sawtooth oscillations with a frequency of 20 cps are produced at its plate and used to counteract the sawtooth action of the search tube. The result is that the local-oscillator frequency hunts around the cavity frequency. This type of circuit, using a reference cavity of $Q \approx 1000$, will stabilize the frequency of a 2K28 10-cm local oscillator to $\pm 0.5$ Mc/sec.

## WIDE-COVERAGE RECEIVERS: CRYSTAL-VIDEO RECEIVERS

BY E. R. GAERTTNER

**8·17. Video Crystals and Crystal Detection.**—The crystal-video receiver consists, as its name implies, of a crystal that detects the r-f signal received from the antenna and an amplifier that amplifies the resulting video signal. For a crystal detector at low levels of signal, the output power is accurately proportional to the square of the r-f signal power. The sensitivity is inherently less than that of a crystal used as a mixer in a superheterodyne, which produces output power proportional to the r-f signal power.

*Components of a Crystal-video Receiver.*—In a crystal-video receiver, the r-f pulse passes from the antenna line through a matching transformer to the crystal detector. The rectified pulse appearing across the output terminals of the crystal detector is then amplified by the video amplifier. The ultimate sensitivity depends upon the r-f match, the crystal, and the coupling into the amplifier. Because of the low detection efficiency of the crystal, the r-f circuit and the video circuits are, in effect, decoupled.

*Crystal Holders.*—At 10 cm, the crystal is mounted at the end of a coaxial line with a suitable choke system for bringing out the video pulse. Matching transformers, either tunable or fixed-tuned, are used to match the r-f line to the crystal. Examples of 10-cm holders are shown in Fig. 8·15.

At 3 cm, the detector crystal is mounted either at the end of a coaxial line or across a waveguide. An example of a tunable 3-cm holder for 1N30 crystals is shown in Fig. 8·16. A commonly used fixed-tuned coaxial holder matches over 50 per cent of all 1N31 crystals to within 1 db loss over the 110 Mc/sec band.

*Frequency Coverage.*—The wideband coverage of the crystal-video receiver can be attributed largely to the inherent insensitivity of crystals to changes of frequency; it is limited only by the frequency sensitivity of the r-f components. The r-f crystal circuit, consisting of the holder and the crystal, can, in general, be represented by a low-$Q$ resonant circuit. At 10 cm it has a $Q$ of 5 to 10. At 3 cm the $Q$ is somewhat larger because of the large r-f transformation ratios required by the capacity of the tungsten-silicon contact. The limits of frequency coverage result, of course, from the frequency sensitivity of these matching structures.

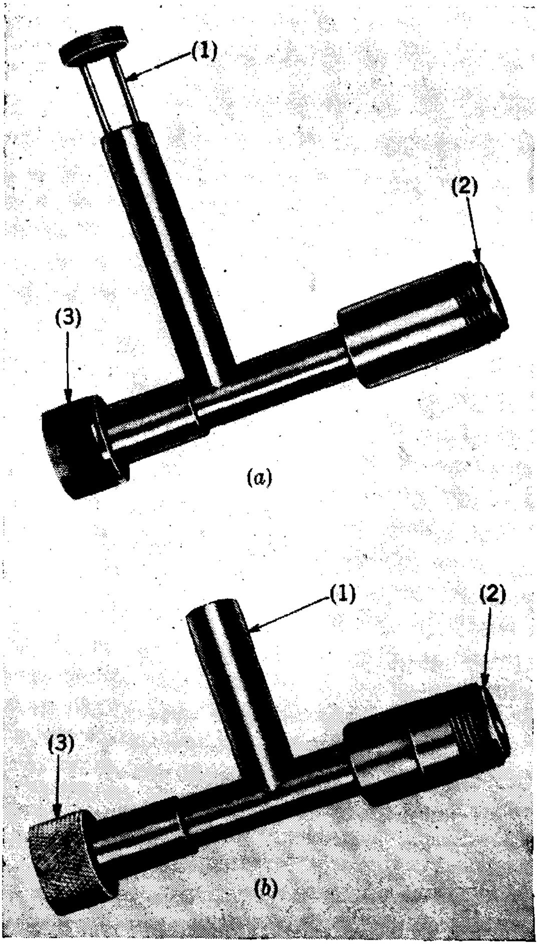


FIG. 8·15.—Ten-centimeter crystal holders. (a) Tunable. (1) tuner; (2) r-f input; (3) video output. (b) Fixed-tuned. (1) fixed tuning stub; (2) r-f input; (3) video output.

*Sensitivity.*—Let us now consider the crystal and its coupling into the video amplifier. Aside from the r-f loss, it is this part of the receiver which determines its ultimate sensitivity.[1]

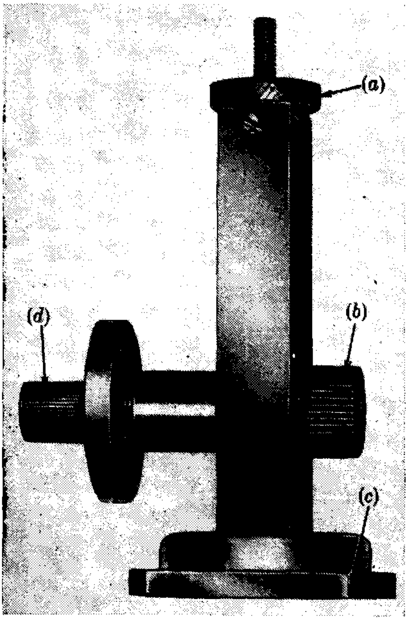


FIG. 8·16.—Tunable 3-cm crystal holder. (*a*) Tuner. (*b*) Cap, removable for crystal replacement. (*c*) R-f input. (*d*) Video output connector.

The process of rectification, that is, the conversion of r-f power into d-c power, is customarily described by the equation

$$i = \frac{P}{b},$$

where $P$ is the r-f power, $i$ is the current into a short circuit, and $b$ is the constant of rectification in watts per ampere and is of the order of magnitude of unity. The crystal can now be regarded as a current source shunted by its video resistance and the combined capacitance of the r-f circuit and the first amplifier tube, as shown in Fig. 8·17. Disregarding the effect of the capacitance on transient signals, the signal voltage appearing on the grid of the first tube is therefore

$$e = R\frac{P}{b}. \tag{2}$$

---

[1] A complete analysis of crystal detection is given in *Crystal Rectifiers*, Vol. 15, Radiation Laboratory Series.

This signal appearing on the grid of the first amplifier tube must, of course, compete with the thermal (Johnson) noise, which originates mainly in the crystal, and the shot and partition noise of the first amplifier tube. If $R_A$ is the equivalent noise-generating resistance of the first stage (approximately 1200 ohms) and $R$ is the video resistance of the crystal, then the root-mean-square noise voltage $n$ generated in these resistances in series is

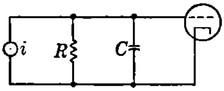


FIG. 8·17.—Equivalent circuit for a crystal detector.

$$n = \sqrt{4kTB(R + R_A)}. \tag{3}$$

$T$ is the absolute temperature, $k$ is the Boltzmann constant, $1.38 \times 10^{-23}$ joule/degree, and $B$ is the video bandwidth. The resistance $R$ can vary from a few thousand ohms to about 20,000 ohms (see Table 8·2) and $R_A$ is about 1200 ohms.

The ratio of Eqs. (2) and (3) gives the signal-to-noise ratio,

$$\frac{e}{n} = \left(\frac{R}{b\sqrt{R + R_A}}\right)\frac{P}{\sqrt{4kTB}}. \tag{4}$$

The bracketed quantity (with $R_A$ taken as 1200 ohms) is largely characteristic of the crystal and is defined as the video crystal's "figure of merit" ($F$). It is seen that for a given receiver bandwidth and specified signal-to-noise-ratio the receiver sensitivity varies inversely as the figure of merit. Values of receiver sensitivity as a function of figure of merit and receiver bandwidth are given in Fig. 8·18.

It is noteworthy that the minimum detectable power varies directly as the square root of the video bandwidth. This means, in general, that increased receiver sensitivity is obtained at the expense of fidelity.

It is also worth remarking that, for a given value of $b$, the higher figures of merit are obtained for higher values of the video impedance $R$. The range of values of video impedance that can be tolerated in any particular application depends upon the fidelity required of the receiver and especially upon whether or not pulse stretching will be objectionable. High-impedance crystals result in a longer time constant at the first grid and, therefore, in longer discharge times of the grid capacity. In receivers designed for pulse-width discrimination, low-impedance crystals must be used.

One effect of changing the video impedance of the crystal is to change the absolute noise level. Crystals with equal figures of merit but different video impedances will have equal signal-to-noise performance but different noise levels. Allowance must be made for this in video amplifier design; a gain control may be necessary.

*Crystal Burnout.*—In some systems the r-f power input can reach large values and the crystal must be able to withstand high pulse powers without damage. It can happen that a ground beacon operating close to a high-powered radar installation will receive signals of 1 watt r-f pulse power. Some approximate numerical values of burnout power are given in Table 8·2.

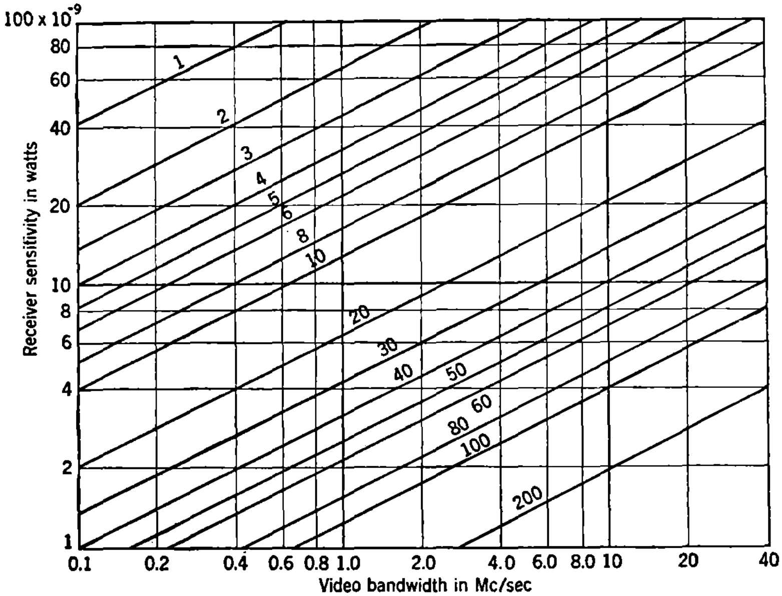


FIG. 8·18.—Sensitivity of crystal-video receiver (signal power = noise power) as a function of video bandwidth for values of the crystal figure of merit shown on the curves. These values are higher than those realized in beacon receivers, where a margin of 4 db or more is required for triggering.

*Video-crystal Specifications.*—The specifications of figure of merit ($F$) and video impedance for a number of commonly used video crystals are summarized in Table 8·2. The values are for an ambient temperature of 20 to 25°C.

TABLE 8·2.—CHARACTERISTICS OF VIDEO CRYSTAL DETECTORS

| Crystal | Band, cm | Figure of merit, minimum | Video impedance, ohms | Burnout, peak watts |
|---|---|---|---|---|
| 1N27 | 10 | 60 | 4000 (max) | 2.5 |
| 1N30 | 3 | 55 | 7000–21,000 | 1 |
| 1N31 | 3 | 55 | 6000–23,000 | 0.5 |
| 1N32 | 10 | 100 | 5000–20,000 | 2.5 |

*Low-temperature Performance of Video Crystals.*—The theory of the rectification process in a crystal leads one to expect that its properties will vary radically with temperature; this is verified experimentally. At low temperatures, the video impedance $R$ becomes very large, and the rectification constant $b$ increases. This tends to maintain the figure of merit constant, but the noise level of the crystal changes with the video impedance. Thus, a set that has been adjusted to trigger perhaps once a second on noise at 20°C may trigger continuously at −10°C.

There is no solution for this within the crystal; it is an inherent property of the device. The practical solution may lie in adjusting the gain as the temperature varies, if the change in video impedance is not otherwise harmful. Alternatively, the temperature range of the crystal may be suitably restricted by temperature control of the crystal holder. The latter alternative is generally preferable. The degree of control required is not great. It is sufficient to keep the temperature of the crystal from dropping below 0°C.

**8·18. Video-amplifier Design.**—The amplifier for a crystal-video receiver[1] must have high gain and also must handle large dynamic ranges for incoming signals. Furthermore, for some applications, high fidelity is required.

*Gain.*—To obtain maximum sensitivity the amplifier should have gain enough to bring pulses just above the noise level up to an amplitude of about 10 volts, sufficient for triggering the succeeding beacon circuits. If one assumes an impedance of the source of 3000 ohms, a 1-Mc/sec bandwidth, and a 3-db noise factor, the noise voltage input at room temperature is 10 $\mu$v. To obtain an output of 10 volts, a gain of about 120 db is required. The gain required will be different for other bandwidths, output levels, and source impedances, but the order of magnitude will remain the same.

A consequence of high video gain is trouble with microphonics, which are not usually annoying in the low-gain video amplifier of a superheterodyne receiver.

Since microphonics are due to mechanical vibrations of the tube elements, their spectrum lies largely within the audio-frequency range. They can, therefore, be eliminated or much reduced in intensity by reducing the gain at these relatively low frequencies. Unfortunately, this cannot always be done without interfering with the fidelity required to give the good reproduction of the pulse shape required for pulse-width discrimination. A compromise must, therefore, be made between excessive microphonics and high fidelity, by introducing a suitable low-frequency rejection circuit into the video amplifier.

---

[1] For a detailed discussion, see *Vacuum Tube Amplifiers*, Vol. 18, and *Microwave Receivers*, Vol. 23, Radiation Laboratory Series.

*Dynamic Range of R-f Signal Strength.*—The overload properties of the amplifier must be good because of the wide range of r-f signal strength encountered. For example, the r-f power received by the antenna can vary from the minimum detectable value of about $10^{-8}$ watt up to 1 watt for a near-by interrogator. Special care must be taken, therefore, to prevent blocking of the amplifier by strong signals for any considerable period of time.

Pulse stretching occurs with the stronger input signals. For an amplifier intended for use with a pulse-width discriminator, it can happen that strong 1-μsec pulses are stretched to 2-μsec pulses in the receiver circuits. These, of course, get through the discriminator adjusted to accept 2-μsec pulses. By choosing video crystals of low impedance and by careful amplifier design, it has been possible to exclude 1-μsec r-f pulses that have an r-f pulse power 60 db above the minimum detectable level for a 2-μsec pulse.

Crystal-video receivers may be classified in many ways, depending upon their applications. For the sake of illustration, they may be classified as follows: high-fidelity receivers and low power-drain receivers. High fidelity is a prerequisite for a receiver suitable for pulse-width discrimination, and low power drain is a prerequisite for a receiver to be used in a lightweight beacon. These two properties are not independent, for, in general, to get a large bandwidth, large plate currents (and small plate load resistors) must be used to obtain adequate gain; this in turn calls for high power drain. Examples of a receiver designed primarily for high fidelity, and one designed primarily for minimum power consumption, are given below.

*High-fidelity Receiver.*—A high-fidelity receiver is used in the AN/CPN-8 beacon, a medium-power 10-cm beacon for ground installations. The specifications call for a minimum triggering sensitivity of 74 dbw and pulse-width discrimination.

The components consist of a 1N27 crystal in a tunable 10-cm holder, and a six-stage resistance-coupled amplifier, using three 6AC7 tubes in the stages where the control grid is driven negative and three 6AG7 tubes in the stages where the grid is driven positive.

The characteristics of the amplifier are the following: gain 120 db, discrimination 50 db, recovery time 25 μsec, bandwidth 1.7 Mc/sec. Aside from amplifier design, stretching is avoided by using low-impedance crystals, less than 4000 ohms. To shape the pulse properly, the video pass band extends down to 100 kc/sec. The gain is still sufficient at audio frequencies to result in some microphonics. These can be overcome for the most part by shock-mounting the first amplifier tube. The recovery characteristic permits the receiver to pass a weak 2-μsec pulse 25 μsec after receiving a strong 1-μsec pulse.

*Receiver with Low Power Drain.*—Such a receiver is used in the AN/UPN-4, a portable 3-cm beacon. Two basic design parameters are lightweight and low drain, because the equipment is battery-operated. The specifications call for a minimum triggering sensitivity of 77 dbw and no pulse-width discrimination.

The components consist of a 1N31 crystal mounted in a 3-cm holder, and a five-stage resistance-coupled amplifier using 3A5 miniature dual triodes.

The characteristics of the amplifier are: gain 120 db, bandwidth 250 kc/sec, and recovery time 150 $\mu$sec. The gain in the audio range is too small for microphonics to give trouble. Multiple overshoots are produced and are "gated" out by the trigger circuit that follows the receiver. The total power drain is 3.2 watts of plate power and 0.9 watt of filament power.

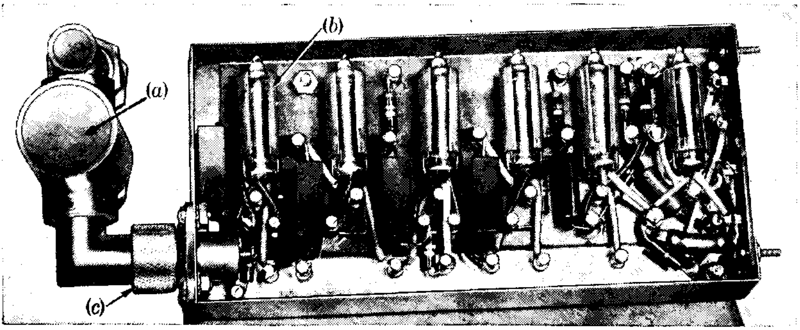


FIG. 8·19.—A 10-cm crystal-video receiver using SD-834 tubes. (a) R-f filter. (b) Tube holder. (c) Crystal holder. Photo is two-thirds actual size.

Figure 8·19 shows a small receiver using SD-834 tubes; it is slightly larger than a cigarette case.

*Miscellaneous Properties.*—Other characteristics of receivers must be considered for certain applications, as, for instance, operation from 28-volt d-c aircraft supplies, special output circuits, two-channel interrogation, and so on. Noteworthy also are types of amplifier coupling other than resistance coupling, such as inductance coupling and direct coupling. For further discussion, see *Vacuum Tube Amplifiers,* Vol. 18, Chap. 4, and *Microwave Receivers,* Vol. 23, Chap, 20, of this series.

# CHAPTER 9

# INTERROGATION CODING

BY C. L. LONGMIRE AND G. P. WACHTELL

It is frequently desirable to use coded interrogation in a beacon-interrogator system to allow the beacon to ignore ordinary radar search signals and concentrate on answering interrogators that desire to "see" it. Interrogation coding is not used in some lightweight, low-power beacons, however, because the decoding circuit adds to the weight and to the power requirements. It also places more stringent requirements on the receiver; this is especially true in pulse-width coding.

The signal characteristics most often used for interrogation coding are radio frequency, pulse width, pulse-repetition frequency, and pulse spacing in multiple-pulse codes. Radio-frequency decoding is introduced quite naturally by using a receiver with a limited bandwidth. Methods of decoding other types of codes are discussed at length in the following sections.

Coded interrogation has uses other than the identification of interrogators desiring to interrogate beacons. One such use, for example, is in remote control of equipment located at the beacon. A discussion of methods for its use in this way is to be found in Chap. 11.

## PULSE-WIDTH DISCRIMINATORS

BY C. L. LONGMIRE

In considering methods of interrogation coding to be used by airborne radar sets, it was early decided that pulse-width coding was the simplest and most economical. Early microwave radars used search pulses of 1-μsec duration or less; it was natural, therefore, to choose 2-μsec pulses as the standard beacon-interrogation signal. The beacon discriminator then had to reject signals of less than 1.9-μsec duration (to allow a margin of safety). Later, when microwave radars with 5-μsec search pulses were designed, the discriminators had to be designed to reject signals longer than about 4.5-μsec as well.

There are two basic circuit types that can be used in pulse-width discrimination; each may have several variations. One may be called the "integrator" type and the other the "delay-line" type.

**9·1. Integrator Discriminators.**—In the integrator type of circuit, the received signal is converted into a sawtooth voltage, of constant slope for all signals greater than the threshold level, and with a duration equal to that of the signal. Obviously, the maximum amplitude of the sawtooth voltage is then directly proportional to the duration of the signal. The sawtooth wave is applied to biased amplifiers arranged to give an output trigger when the peak of the sawtooth voltage is within the proper limits.

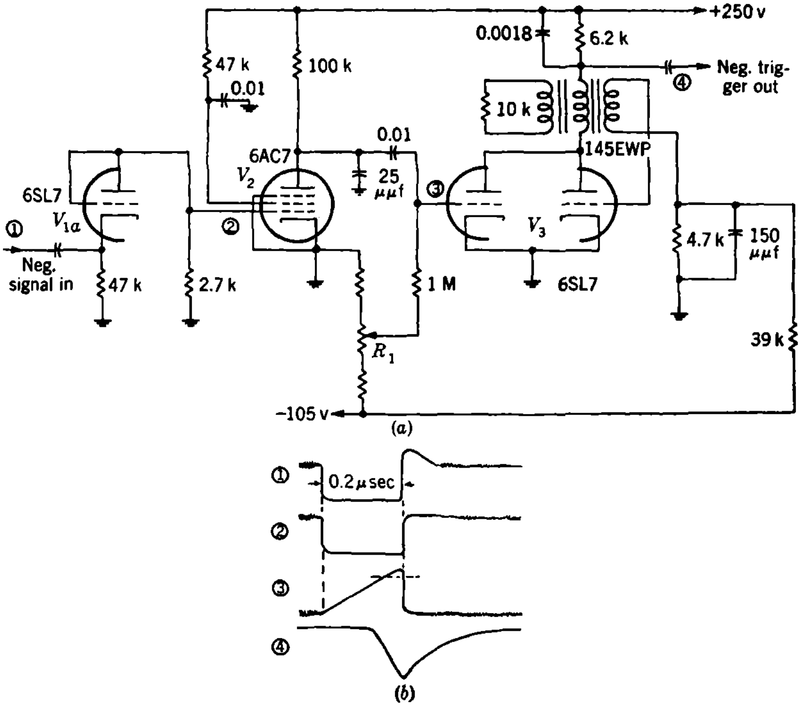


Fig. 9·1.—(a) An integrator-type discriminator rejecting short pulses. The minimum pulse width accepted is determined by a potentiometer adjustment. $V_1$—input diode. $V_2$—integrator. $V_3$—biased amplifier. $R_1$—potentiometer for adjustment of minimum pulse width accepted. (b) Waveforms at numbered points.

*Discriminator to Reject Short Pulses.*—An example of an integrator-type discriminator designed to reject short pulses only is shown in Fig. 9·1. Waveforms are sketched for several points. The signal from the receiver is negative and limits at 20 to 40 volts. On the end of the signal there is apt to be a positive "overshoot" of short duration, and the baseline is cluttered with noise, 6 to 10 volts peak to peak.

The signal is first passed through a diode to remove the positive overshoot. If this were not done, the overshoot would cause grid current to flow in the integrator tube that follows; this current would produce an additional voltage drop across the coupling condenser, which would persist for some time after the positive overshoot ended. The resulting shift in integrator grid voltage would be greatly amplified in the integrator plate circuit, and the sawtooth waveform for a following signal would start from a different voltage level, causing a shift in the minimum acceptable pulse width. The diode also removes the positive part of the noise and, by virtue of a small bias built up on the coupling condenser by rectification of noise, some of the negative part.

The signal, still negative, but with the overshoot and some noise removed, is then applied to the grid of the integrator tube. The grid return resistance must be fairly low because it must discharge the stray capacitance at the end of a signal when the diode does not conduct. The integrator is a sharp cutoff pentode normally at zero bias, the plate circuit of which consists of a parallel resistance-capacitance circuit with a time constant of about 5 μsec, including the effect of stray capacitance. When the negative signal is applied to the grid, plate current becomes zero for all signals above the threshold level, and the plate voltage rises with an exponential waveform that is approximately linear for the first 2 or 3 μsec. At the end of the signal, plate current resumes, and the plate-circuit capacitance is discharged rapidly to the original voltage level. Quick return of the plate voltage is desirable, because it reduces additive effects for closely spaced signals and thereby reduces plate-voltage fluctuations resulting from noise.

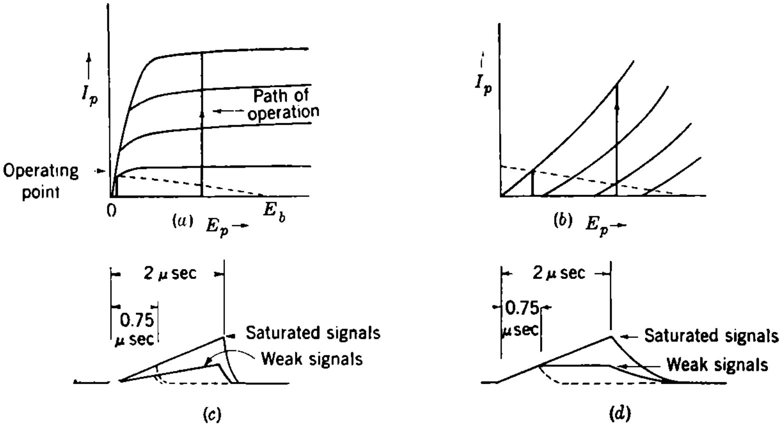


FIG. 9·2.—Characteristics of pentode and triode integrators. In (a) and (b) are shown the plate characteristics of pentodes and triodes for various values of grid bias; in (c) and (d) are shown the corresponding integrator waveforms. In (c) and (d) the full lines show the operation with 2-μsec signals and the dotted lines with 0.75-μsec signals.

To achieve a quick return, the plate-circuit capacitance should be small and the tube should be capable of drawing large currents. The minimum capacitance is the stray capacitance, but this should not be used alone as it may vary from set to set.

It is usually possible to obtain larger currents for low plate voltages by using pentodes rather than triodes with the same cutoff voltage [see Fig. 9·2 (*a*) and (*b*)]. Another advantage of the pentode is the removal of some baseline noise by virtue of the coincidence of several grid-voltage curves at the operating point. A third advantage is that signals weaker than the threshold level do not have so great an effect as in a triode circuit [see (*c*) and (*d*) of Fig. 9·2]. Weak short signals (noise) are "amplified" more in the triode circuit. Noise on the integrator plate results in "time jitter" and broadens the discrimination.

The sawtooth voltage is applied to the grid of the biased amplifier, the bias of which is adjusted so that, when the signal is longer than the minimum acceptable signal, the tube conducts and delivers a trigger to the blocking oscillator, which is normally biased off. The discriminator output trigger is obtained from the plate circuit of the blocking oscillator. The recovery time of the blocking oscillator is kept small by using a bias network with a small time constant. Quick recovery in all discriminator circuits is essential to maintaining a constant level of discrimination.

It is clear that the delay in this discriminator is constant, barring threshold effects, with respect to changes in signal width and amplitude. Threshold effects are discussed below (Sec. 9·6). The actual delay from leading edge of signal to leading edge of output trigger is about 2.5 μsec when the circuit is adjusted to discriminate at 1.9 μsec. Of the total delay, 1.9 μsec is unavoidable, because it is impossible to tell whether a signal is 1.9 μsec long or longer without allowing the signal to continue for at least that length of time. The remainder of the delay is due to the blocking oscillator.

*Discriminator to Reject Short and Long Pulses.*—The integrator principle can also be used for rejection of long signals. Figure 9·3 shows an example of a "trailing edge" type of discriminator, so called because the output trigger is initiated by the end of the signal. It is evident that, for the discriminator to accept only signals between 2 and 5 μsec in duration, there must be a delay of at least the duration of the signal in question.

In Fig. 9·3, the diode and integrator stages are identical with those of the circuit just discussed. Following the integrator is a biased cathode follower with a differentiating network in the plate circuit. If the sawtooth output of the integrator is large enough to drive the cathode follower to conduction, the differentiating network generates a rather sharp positive trigger at the end of the signal. The bias on the cathode follower is so adjusted that for a signal 1.9 μsec long this trigger is just large enough to initiate regeneration in the blocking oscillator, which then produces the output trigger.

The cathode circuit of the cathode follower reproduces that part of the sawtooth input which is above the conduction level and passes it on to the grid of the biased amplifier. The bias of this tube is adjusted so that signals longer than 4.5 μsec cause conduction; the negative waveform generated in the plate circuit then cancels the positive trigger, which comes from the differentiating network at the end of the signal, and the blocking oscillator is not driven into the regenerative region.

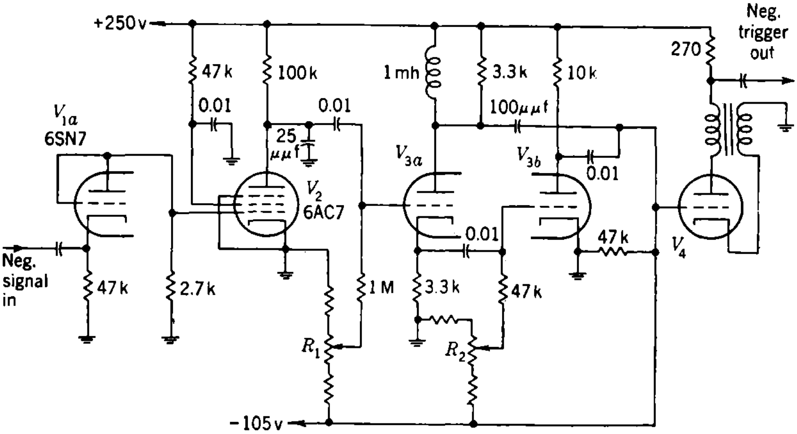


FIG. 9·3.—An integrator-discriminator which accepts only pulses between 2 and 5 μsec long. $V_1$—diode; $V_2$—integrator; $V_{3a}$—differentiator and cathode follower; $V_{3b}$—biased amplifier; $V_4$—blocking oscillator; $R_1$—adjustment for minimum pulse width accepted; $R_2$—adjustment for pulse-width interval accepted.

There are doubtless many possible variations of the integrator type of pulse-width discriminator. The examples given are discriminators that have been used in beacon circuits and have proved satisfactory.

**9·2. Delay-line Discriminators.**—Unlike the integrator type of discriminator, which may be said first to convert pulse width functionally into pulse amplitude and then to discriminate on the basis of pulse amplitude, the delay-line discriminator actually discriminates on the basis of pulse width. The principle used, in general, is to compare the time duration of the signal with a fixed standard time interval obtained from a delay line or other delay device.

*Discriminator to Reject Short Pulses.*—Figure 9·4 shows an example of a delay-line discriminator for rejection of short pulses. The purpose of the diode is, as in the integrator discriminator, to remove positive overshoots and to eliminate a substantial portion of the receiver noise. The next stage is a limiter that saturates at about 4 volts; signals below this level are disregarded. The delay line is matched in impedance at the input end and open-circuited at the other. A voltage step applied at the input at a given instant travels down the line, is reflected, appears at the input 1.6 μsec later, and adds to the initial voltage step. No further reflections occur because of the impedance match.

If a 1-μsec signal is applied, the waveform at the amplifier grid is as shown in Fig. 9·4. The bias of the amplifier tube is large enough so that this waveform does not cause conduction (the waveform is of constant amplitude for all limited signals). However, if a 2 -μsec signal is applied, the initial pulse and the reflected pulse overlap in time at the amplifier grid, which is driven twice as far positive as before. The amplifier then conducts and triggers the blocking oscillator, which delivers the output trigger.

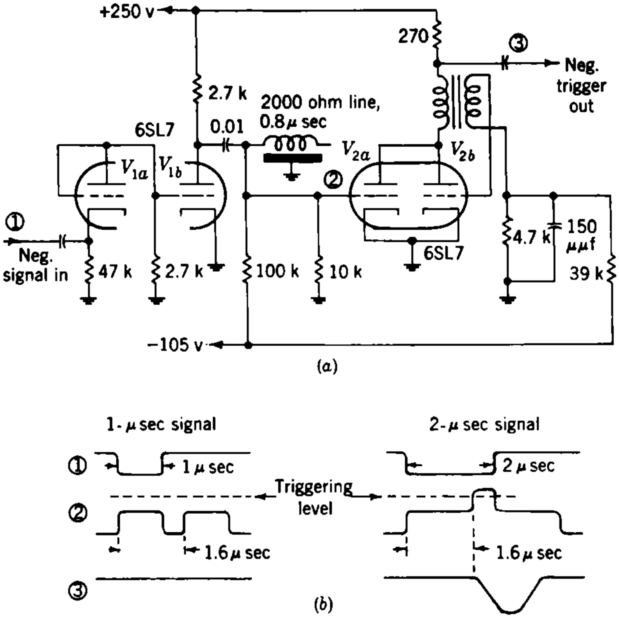


FIG. 9·4.—(a) A delay-line discriminator which rejects short pulses (less than 1.6 μsec in this case). The minimum pulse width accepted is determined mainly by the delay line. $V_1$—diode; $V_{1b}$—limiter; $V_{2a}$—biased amplifier; $V_{2b}$—blocking oscillator. (b) Waveforms at numbered points in (a), for both 1-μsec signals (which are rejected) and 2-μsec signals (which are accepted).

The delay in this discriminator must be longer than the time required for a signal to travel down the line and back. The total delay from leading edge of signal to leading edge of output trigger is about 2.5 μsec, including the time required by the blocking oscillator. Barring threshold effects, the delay is constant for changes in signal length and amplitude.

*Discriminator to Reject Long and Short Pulses.*—Figure 9·5 shows an example of a delay-line discriminator designed to reject short and long pulses. No input diode is shown, although a diode may be necessary if considerable noise and large overshoots exist. The first stage shown, the limiter, should saturate on all usable signals. The delay line coupled to the plate of the limiter is short-circuited at its output end and matched in impedance at its input end by $R_1$. The plate resistance of the limiter tube should be much greater than $R_1$ so that there will be a good match whether the tube is conducting or not. Typical waveforms are shown in Fig. 9·5*b*. The short-circuited delay line transforms every voltage step into a rectangular voltage pulse of the same sign and 0.9 μsec long. Each signal is therefore converted into two rectangular pulses, a positive one starting with the leading edge of the signal, and a negative

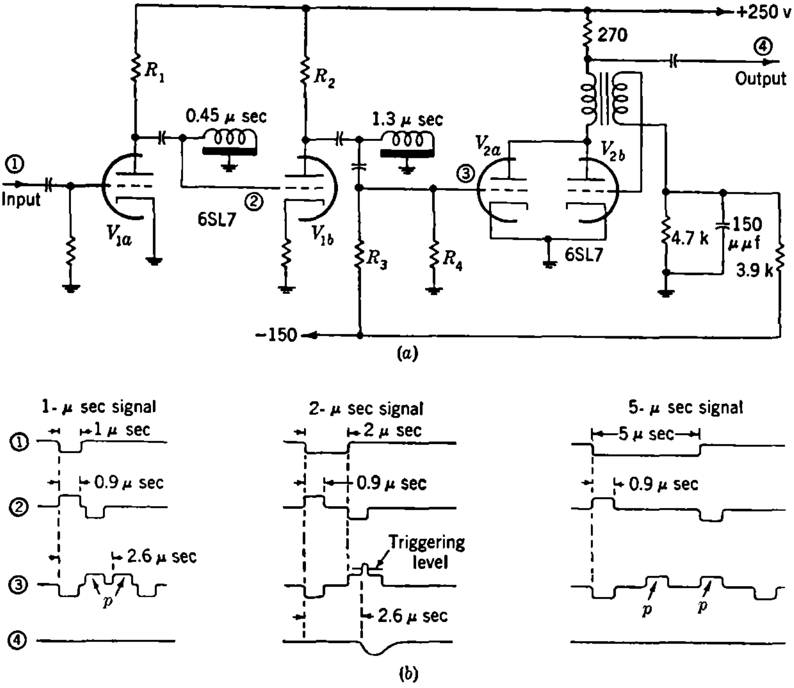


FIG. 9·5.—(*a*) A delay-line discriminator which accepts only pulses between 1.7 and 3.5 μsec wide. The acceptance interval is 2 × 1.3 μsec ± 2 × 0.45 μsec. $V_{1a}$—limiter; $V_{1b}$—buffer-inverter; $V_{2a}$—coincidence amplifier; $V_{2b}$—blocking oscillator. (*b*) Waveforms at numbered points for 1-, 2-, and 5-μsec pulses. *p*—Pulse reflected by 1.3-μsec line.

one starting with the trailing edge of the signal. The buffer-inverter inverts these (and therefore must be able to pass both positive and negative pulses) and passes them into the 1.3-μsec delay line. The pulses travel down this line, are reflected with opposite sign, and appear again at the input 2.6 μsec later. Further reflections do not occur because $R_2$, $R_3$, $R_4$, and the buffer-inverter plate resistance in parallel are chosen to match the line impedance. Therefore, at the grid of the coincidence amplifier there are four successive 0.9-μsec rectangular pulses, the first and last of which are negative whereas the intervening two are positive. The bias of the coincidence amplifier is such that neither of the positive pulses alone will cause conduction; but if the two positive pulses overlap in time, the grid of the coincidence amplifier is driven twice as far in the positive direction as before; the coincidence amplifier then conducts and triggers the blocking oscillator. The two positive pulses overlap enough to cause triggering for signals between 1.9 μsec and 3.3 μsec long. This allows a 0.2-μsec overlap as the minimum amount to cause triggering.

In Fig. 9·5, the limiter and buffer-inverter are shown as triodes. In some cases, it may be necessary to use pentodes to preserve pulse shape and still have sufficient gain. Delay lines that have good frequency characteristics should be used because less sharp discrimination results from rounding off the edges of pulses.

The delay in this discriminator, barring threshold effects, is constant at slightly more than 2.6 μsec for signals between 1.9 μsec and 2.6 μsec long. For signals between 2.6 μsec and 3.3 μsec long, the delay is slightly greater than the duration of the signal.

There are many other variations of the delay-line type of circuit. Different types of coincidence circuits and different arrangements of delay lines can be used. The circuits described have the advantage of being fairly economical with respect to delay lines, because pulses are run down and back to get the same delay with half the usual amount of line.

**9·3. Comparison of Integrator and Delay-line Discriminators.**—It is an experimental result that the integrator type of discriminator can be adjusted to give a sensitivity about 3 db better than the delay-line type for the same amount of noise firing. The integrator type can be set to accept tangential signals (see Sec. 8·3) without triggering appreciably on noise alone, but the delay-line type, if set to accept the same signal, triggers excessively on noise. The advantage is inherent in the integration. In the delay-line type, a high, narrow noise "spike" may appear at the coincidence tube at just the right time to coincide with a 1-μsec signal and cause an output pulse. In the integrator type the same kind of spike would, at worst, only make a 1-μsec signal look a little longer. In other words, the video bandwidth of the integrator is narrower than that of the delay-line type; noise is correspondingly diminished.

The integrator type is more easily adjustable over a wide range than is the delay-line type. On the other hand, it is necessary to adjust the integrator type but not the delay-line type, which may be factory-set at the proper value.

Delay lines at present are fairly bulky and usually present a special problem to a manufacturer. The integrator type uses, in general, no more tubes than the delay-line type.

In integrator-type discriminators that must reject signals longer than beacon-interrogation pulses, weak, long signals may cause triggering. If the integrator tube is not cut off, the sawtooth voltage has a smaller slope and may not reach the amplitude necessary to cause rejection. For still weaker signals, the sawtooth voltage does not even reach the amplitude corresponding to a 2-μsec signal. It is apparent that the range of signal amplitude in which erratic triggering results is proportional to the acceptance interval of the discriminator (see Fig. 9·6). If the acceptance interval is from 2 to 3 μsec, the ratio of maximum-to-minimum signal amplitude for erratic triggering is $\frac{3}{2}$,—1.8 db for the square-law crystal-video receiver and 3.6 db for the superheterodyne receiver. Noise would increase each of these values slightly, possibly to 3 and 5 db, respectively. Now, 5 db corresponds to a factor of 1.8 in range, according to the inverse square law. Thus, in the worst case, all long-pulse search radars of a given type between 100 and 180 miles from the beacon might trigger a beacon with a superheterodyne receiver of the appropriate sensitivity.

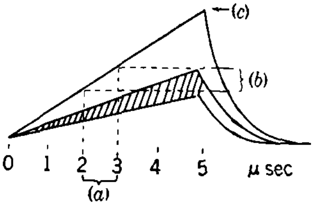


FIG. 9·6.—The effect of weak, long pulses in a discriminator designed to reject long pulses. If the discriminator accepts pulses between 2 and 3 μsec long, weak 5-μsec pulses which do not saturate the integrator grid may be accepted. (a) Acceptance interval in signal length. (b) Acceptance interval in amplitude. (c) Saturated signal.

Obviously, the seriousness of the defect depends upon the use of the long-pulse search radars and the sensitivity of the receiver. The latter is unlikely to have just the critical value to cause failure of satisfactory discrimination at horizon range. The defect can be removed, at the expense of adding tubes, by adding an amplitude discriminator that will not allow the pulse-width discriminator to trigger the coder unless the input signal is large enough to cut off the integrator. Circuits of this kind are described in the following section.

From these considerations, it is apparent that the delay-line type of pulse-width discriminator should be used when noise is small compared with the minimum signal and a system that requires no adjustment is desired. The integrator type has more to offer, on the other hand, for adjustability, ease of manufacture, and sensitivity.

**9·4. Delay Considerations.**—When it is desired to range on a beacon with great accuracy (to about 20 yds, for example), delay in the beacon cannot be neglected. It is easy to compensate for the beacon delay in the range calculations, provided the value of the beacon delay is known. This implies that the beacon delay (and therefore the discriminator delay) must be maintained constant to ±0.1 μsec over reasonably long periods of time for the above range accuracy to be attainable. Delay accuracy to ±0.1 μsec except for weak signals can be had with either type of discriminator, although the integrator type may, perhaps, require more frequent checking.

Many factors tend to cause variations in discriminator delay. Corrections for some of these, like tube aging and replacement of tubes or components, can be made during periodic checks by adjustment of the over-all delay. Delay variations due to other causes, such as changes in interrogation-pulse width and amplitude, receiver noise, voltage-supply variations, and temperature variations, must be minimized by suitable circuit design.

When serious delay variations are caused by temperature changes, they can usually be eliminated by using capacitors with the correct temperature coefficient in the delay network of the discriminator. For instance, if increasing temperature causes the resistance in the plate circuit of the integrator tube to increase, the time constant of the *RC*-circuit can still be maintained constant by using a capacitor with a negative temperature coefficient of the correct magnitude.

The effects of supply-voltage variations are usually made negligible by using a regulated supply. There are, however, methods of compensation for voltage variation. One is to let a bias voltage change in such a way as to counteract the effect of change of plate voltage.

Noise spikes add to or subtract from the signal, when signals close to noise are used. The leading edge then appears earlier or later than it should. The only successful method of removing this kind of time jitter is to increase the ratio of signal to noise.

Other methods proposed are based on the idea of averaging out the noise by viewing several pulse-repetition cycles at once; they require a knowledge of the repetition rate of the interrogating signal. Systems of this type using supersonic delay lines or storage tubes have been developed. Since several scanning radars with different pulse-repetition frequencies are often interrogating a single beacon, which must answer each radar, pulse for pulse, this idea does not seem practical for beacons.

*Effect of Signal Length.*—Changes of discriminator delay with signal length are of two types: threshold variations, which occur when the signal is barely long enough to trigger the discriminator, and inherent variations, which are fundamental in the method of discriminating.

The final stages of all the discriminators shown in the two proceding sections are regenerative trigger circuits, which, for practical purposes, have the property of putting out either a full-sized pulse or nothing, depending on the size of the trigger from the preceding stage. Now, if the signal length is varied continuously across the threshold value, the size of the trigger to the trigger circuit changes continuously from a value too small to a value sufficient to initiate regeneration. The action of regenerative circuits is not instantaneous, because of the retarding effects of shunt capacitance and (in this case) leakage inductance. The action is an accelerating change somewhat of the form $Ae^{t/\tau}$; larger triggers may be considered as resulting in larger values of the constant $A$. As a result, the output pulse rises to saturation in a shorter time than for smaller triggers. As $A$ approaches zero (corresponding to the threshold), the rise time, and therefore the discriminator delay, may increase without limit. Actually, we have oversimplified, for the regenerative circuit has a minimum frequency for which the gain of the circuit is greater than unity, so that, whatever waveform results, the voltage must return to the initial value within a half period of this minimum frequency.

The question "Why not use a high-gain amplifier instead of a regenerative circuit?" might be asked. The answer is that usually somewhere in the beacon there is a trigger circuit; wherever the first one is, there the same variation in delay will occur. Even if high-gain amplifiers are used, delay variations at threshold will still occur.

For the blocking oscillators used, the maximum shift of output trigger at the threshold is about a microsecond. In the circuit of Fig. 9·1, Sec. 9·1, the region of pulse width over which objectionable threshold variations occur is about 0.2 μsec. This region can be reduced in size by using more gain before the blocking oscillator; it is usually sufficient, however, when constant delay is desired, to set the discriminator slightly lower than normal so that 2-μsec pulses will be clear of the threshold region.

The discriminator shown in Fig. 9·3, Sec. 9·1, has inherent delay variations with signal length because the output trigger is initiated by the end of the signal. This type of discriminator is not suitable for accurate ranging, unless special precautions are taken to prevent stretching of the signal in the receiver. The discriminators of Figs. 9·1 and 9·4 have delays inherently independent of signal length. The circuit of Fig. 9·5, Sec. 9·2, has a delay inherently independent of signal length for signals between 1.9 and 2.6 μsec.

**9·5. Constant-delay Discriminator for both Long and Short Pulses.**—A discriminator of the integrator type to reject short and long pulses can be designed to have a delay inherently independent of signal length; a suggested block diagram is shown in Fig. 9·7. The biased amplifier forms a negative rectangular pulse from the portion of the sawtooth voltage that comes 2 μsec after the beginning. This rectangular pulse travels down a 1-μsec delay line which is terminated in a small condenser that short-circuits only the high frequencies in the pulse. The reflected wave, which is absorbed at the input end, is shown by itself in Fig. 9·7; a short positive trigger occurs at the input end 4 μsec after the beginning of the signal. If the signal is still present, the positive trigger is canceled by the negative pulse still present at the plate of the biased amplifier. If the signal has ended at 4 μsec, the positive trigger drives the coincidence tube to conduction; this tube in turn triggers the blocking oscillator. The output trigger always comes at slightly after 4 μsec. In this circuit, special precaution must be taken to maintain the d-c level at the biased amplifier grid, because grid current flows regularly; either d-c coupling from the integrator or d-c restoring can be used. The bias on the biased

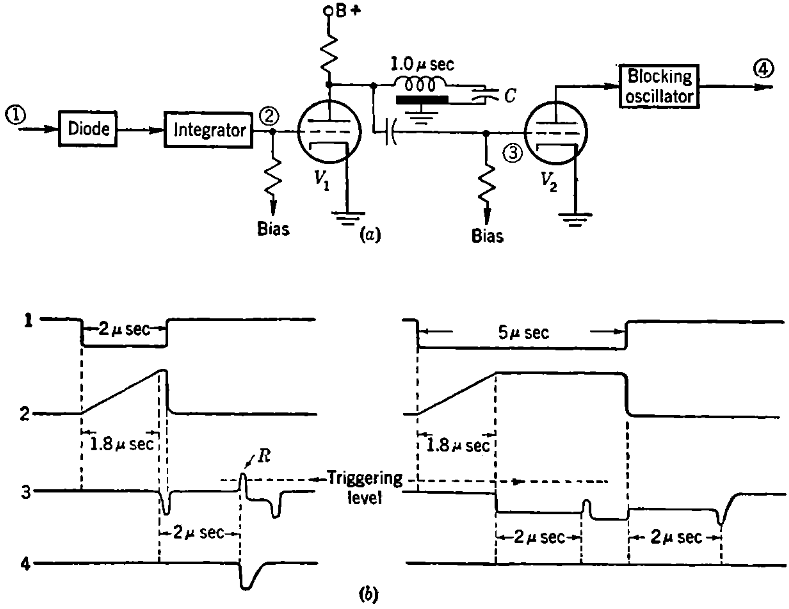


FIG. 9·7.—(a) An integrator type of discriminator with constant delay, which accepts 1.8- to 3.8-μsec pulses but rejects shorter and longer pulses. C is a short circuit for high frequencies. $V_1$—biased amplifier; $V_2$—amplifier. (b) Waveforms at numbered points for 2- and 5-μsec signals. R—reflected pulse.

amplifier determines the lower level of acceptance, and the delay line fixes the interval of acceptance.

**9·6. Effect of Signal Amplitude.**—The discriminator delay varies considerably for signal amplitude below some particular level. Signals that do not cut off the integrator tube cause sawtooth waveforms of a lesser slope than normal, and the biased amplifier is turned on at a later time. The delay-line type of discriminator also has this defect; waveforms below the limiting value at the coincidence tube cause a smaller trigger to the blocking oscillator, which therefore triggers with greater delay. For weak signals, the delay is likely to be as much as a microsecond greater than normal.

Ordinarily this weak-signal error is not objectionable. When a radar is scanning, the first and last beacon responses of each scan are the only ones likely to be in error, because for these the interrogation pulses are near the threshold. The intervening responses can be ranged on by an operator without objectionable inaccuracy. If automatic-tracking circuits having "velocity memory"[1] are used, however, an inaccuracy of the last pulse during a scan may affect the remembered velocity seriously; this results in "jumpy" range data. The error depends upon the number of pulses per scan and the constants of the memory circuit.

It may be desired, therefore, to take steps to ensure that the beacon delay is correct even for weak signals, if the beacon responds to them at all. This can be assured by suppressing all responses for which the delay is not correct. (These threshold variations in delay have nothing to do with noise; there is nothing to gain by increasing the ratio of signal to noise.)

Figure 9·8 shows a block diagram of a discriminator that corrects for weak-signal variations. The method used is to place an amplitude discriminator that triggers only on signals above a given amplitude level in parallel with an ordinary discriminator having constant delay for all signals above the same amplitude level. A coincidence circuit is then used to give an output only when both discriminators fire. Note that the amplitude discriminator itself must contain a regenerative circuit, so that its output waveform will also be subject to delay variations for signals that barely trigger it. Therefore, its output waveform must be "flat" over the entire interval within which the trigger from the pulse-width discriminator may arrive. A flat-topped waveform is obtained by using a rectangular pulse generator—the current waveform from a blocking oscillator, for instance. For the weakest signal accepted, the rectangular pulse must reach its maximum before the trigger from the pulse-width discriminator. The signal is differentiated before triggering

---

[1] Circuits that store in a resistance-capacitance network a voltage proportional to the rate of change of range are said to possess "velocity memory."

the rectangular-pulse generator in order to reduce delay variations of the rectangular pulse.

**9·7. General Characteristics of Pulse-width Discrimination.**—The circuits that have been described can discriminate perfectly between 1- and 2-μsec signals, and between 2- and 5-μsec signals. One might suppose, then, that the problem of pulse-width discrimination is completely solved. Such is not the case, however, for there are tendencies in the beacon-interrogator system for the interrogator pulse to be increased in length before arriving at the discriminator. This stretching arises from two major sources: receiver pulse-stretching and echo-stretching.

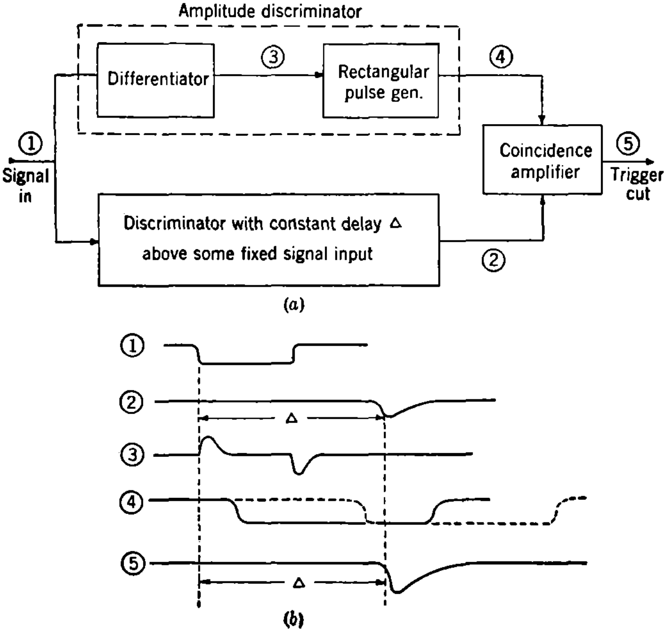


Fig. 9·8.—(a) A discriminator in which the over-all delay Δ is independent of signal strength. The amplitude discriminator prevents responses of incorrect delay. (b) Waveforms at numbered points. In waveform (4), the amplitude-discriminator pulse may move between the limits indicated by the full and dotted lines with varying signal strength.

*Receiver Pulse-stretching.*—Receiver pulse-stretching is mainly the result of overloading in the video stages of the receiver, combined with resistance-capacitance distortion, as shown in Fig. 9·9. In typical high-fidelity crystal-video receivers, increasing the r-f power by 60 db above the threshold value increases the receiver-output signal width by about 1 μsec. Such a receiver is said to possess 60 db of discrimination between 1- and 2-μsec signals. In superheterodyne receivers, when the number of video stages is only one or two, the stretching for a 60-db increase in r-f power is normally less than 0.25 μsec, and this is not objectionable.

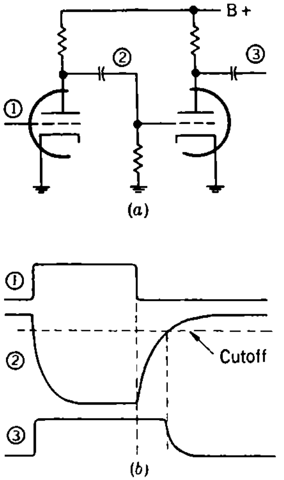


FIG. 9·9.—The origin of pulse stretching in a video amplifier. (a) *RC*-coupled amplifier. (b) Waveforms at numbered points.

*Stretch-amplitude Compensation.*—The discrimination can be improved a few decibels by using the increase in signal amplitude to cancel the stretching associated with the increase in amplitude. Unless changes are made in the receiver, it is possible to cancel only the stretching that occurs between tangential and limiting signals at the receiver output. For the crystal-video receiver, this improves the discrimination by about 12 db (or a factor of 4 in range).

A modification to introduce stretch-amplitude compensation into the integrator type of discriminator is shown in Fig. 9·10. In the diagram, *C* is a blocking condenser, and *R* is chosen so that a negative signal that is *just* limited at the receiver output subtracts from the sawtooth amplitude an amount equivalent to the stretching between tangential and limiting signals. It is then necessary, when adjusting the discriminator, to check the acceptance of 2-μsec signals at both tangential and limiting levels.

By taking a signal from an early stage in the receiver, where there are more decibels between tangential and limiting signals, a greater improvement in discrimination is theoretically possible. It is technically difficult to do this, however. without causing oscillations.

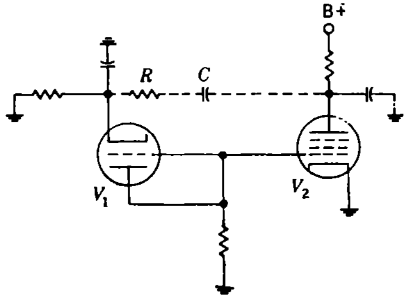


FIG. 9·10.—An *RC* feedback circuit which diminishes pulse stretching in an integrator discriminator by "stretch-amplitude compensation." $V_1$—diode; $V_2$—integrator.

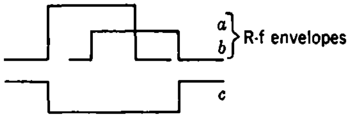


FIG. 9·11.—The origin of echo-stretching. (a) Direct signal. (b) Reflected signal. (c) Receiver output.

*Echo Stretching.*—Echo stretching results when signals reach the beacon antenna by both direct and reflected paths. The reflected signal necessarily arrives slightly later than the direct signal. The receiver may limit even on the reflected signal, however, so that the receiver output signal is as long as the direct and the reflected signal combined. (See Fig. 9·11.)

It is obvious that nothing can be done in the discriminator to remedy this situation. Echo stretching is discussed in Secs. 8·11 and 20·1.

## OTHER TYPES OF DECODER

**9·8. Double-pulse Decoders.**[1] *General Considerations.*—In double-pulse coding the interrogator transmits two pulses a few microseconds apart; normally it would transmit just one pulse. Circuits at the beacon cause it to respond if the separation of the pulses is correct for that beacon. Thus, different beacons can have different interrogation codes.

The problem of double-pulse decoding can be treated with varying emphasis on security, by which we mean freedom from response to incorrect interrogation codes. The simplest and least secure method is for the beacon to accept signals that exist at any two instants separated by a given amount of time $\Delta t$; here a single pulse longer than $\Delta t$ might cause triggering. A more complex method is for the beacon to accept only pairs of pulses, the leading edges of which differ by an amount $\Delta t$, and which, in addition, do not have other pulses between them.[2] Another method is for the beacon to accept any two pulses the leading edges of which are separated by an amount $\Delta t$; in this case a series of closely spaced narrow pulses might cause triggering. This last method is perhaps the best for general use, and it alone will be described here.

The usual method of decoding is to delay the first pulse by an amount equal to the correct value for interrogation; a coincidence detector then gives out a trigger if the second pulse coincides in time with the delayed first pulse. Two methods of obtaining a delay have been considered: multivibrators and delay lines. Multivibrators have a dead time after each signal triggers them; this type of circuit, therefore, has a definite maximum working rate. Delay lines do not have this defect, and they have the advantage of being less subject to unwanted variations. For these two reasons they have found favor. However, in some systems in which there are few interfering signals (signals not intended to interrogate the beacon), the maximum working rate limitation of multivibrators does not become objectionable. The general question of the relative virtues

---

[1] By C. L. Longmire and G. P. Wachtell.

[2] A circuit which prevents a response when an intervening pulse is present is sometimes called a "wonter."

of trigger and storage devices, which is pertinent to this argument, has already been discussed in Chap. 5.

Another factor that must be considered is the amount of delay required. Electronic delay lines at present are practical only for delays up to about 10 μsec, whereas the multivibrator is best suited for delays

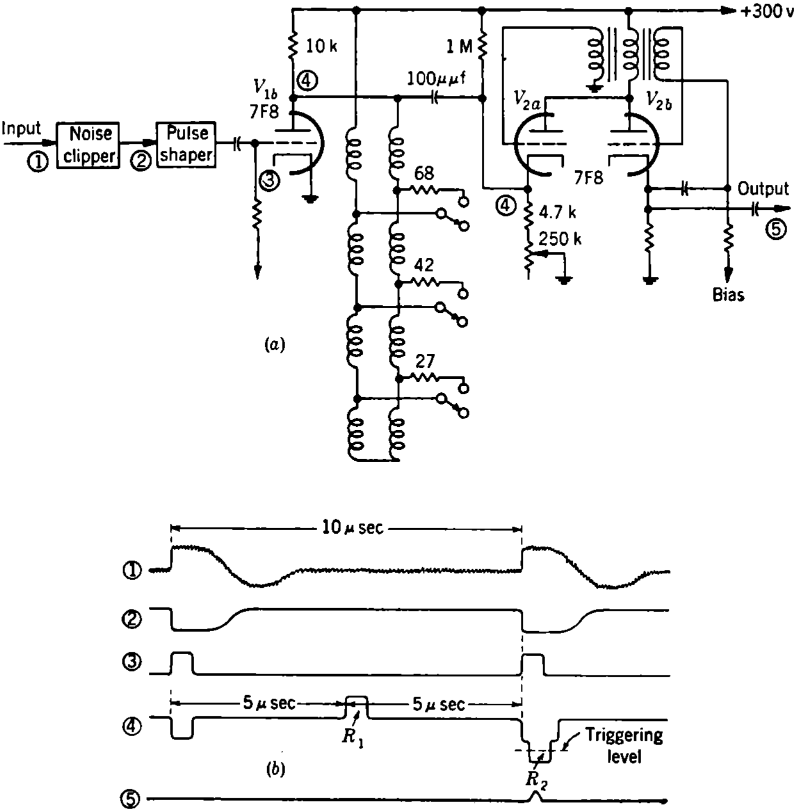


FIG. 9·12.—(a) Double-pulse decoder using delay lines. The delay line has an impedance of 650 ohms and a total delay, with the switches open, of 2.5 μsec. $V_{1b}$—buffer; $V_{2a}$—biased amplifier; $V_{2b}$—blocking oscillator. (b) Waveforms at numbered points in (a). $R_1$ and $R_2$ are the first and second reflections of the initial pulse as they appear at the input terminals of the delay line.

greater than 5 μsec. For code intervals longer than 10 μsec, the advantages of storage devices can be preserved by using supersonic delay lines.

*Delay-line Type.*—One form of double-pulse decoder that has been used successfully is shown in Fig. 9·12. The circuit shown can be made to accept pulses with any one of four values of pulse separation. The acceptable pulse separation is changed by changing the position of the short circuit on the delay line.

The input signal is cleared of noise, shaped, and limited. The resultant pulses are then applied to the grid of the buffer. The first pulse is inverted by the buffer and fed into the delay line as a negative pulse. (The biased amplifier does not conduct.) This pulse travels down the line, is reflected with a change of sign, and returns to the input of the line as a positive pulse. The impedance shunting the input of the line is much higher than the impedance of the line; the pulse is, therefore, reflected again, this time without a change of sign. The pulse then makes another trip down the line and back, finally returning as a negative pulse. If another pulse is applied to the grid at the same instant that this negative pulse reaches the plate of the buffer, the two pulses combine at the plate of the buffer. The resultant pulse is large enough to drive the biased amplifier to conduction and to trigger the blocking oscillator.

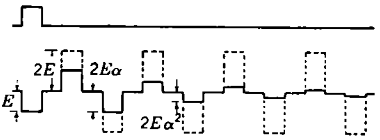


FIG. 9·13.—The response of the delay line in Fig. 9·12 to a single pulse of amplitude $E$. The attenuation of the delay line for two round trips is $\alpha$.

Each pulse is reflected several times in the line. Figure 9·13 shows the response of the line to a single pulse.

If it were not for attenuation, all of the reflected pulses would be twice as large as the initial pulse fed into the line, because the impedance shunting the input terminals of the line is much greater than the characteristic impedance of the line. Obviously, some attenuation is desirable so that coincidences with reflections other than the first negative one will not drive the biased amplifier to conduction. To get the greatest range of safety for the bias on the biased amplifier, it is necessary to choose the attenuation so as to give the greatest possible difference in amplitude between the first and second negative reflections. Mathematically, if $\alpha$ is the attenuation for two round trips on the line, it is necessary to maximize the expression $2E\alpha - 2E\alpha^2$. This is done by making $\alpha$ equal to $\frac{1}{2}$; then, the first negative reflection is equal in amplitude to the initial pulse. The resistors associated with each short-circuiting switch are chosen to give the correct amount of attenuation. Noise and threshold effects in this circuit are similar to those in the delay-line type of pulse-width discriminator, because the circuits are similar.

The second pulse in a double-pulse signal is usually the range-reference pulse of the interrogator; it is, therefore, better to time the decoder output trigger from this pulse rather than from the first. To do this it is necessary to choose the length of the line so that the second pulse of the signal arrives at the plate of the buffer slightly after the first negative reflection; it is then the second pulse that actually drives the biased amplifier to conduction.

*Multivibrator Type.*—Because both multivibrator and delay-line decoders operate on the same general principles, the multivibrator type will not be discussed at great length. A block diagram of this type of discriminator is shown in Fig. 9·14. The first step is to remove noise and shape the signal as in the preceding section; in some systems, however, this may not be necessary. The first signal then triggers a biased multivibrator which produces a rectangular pulse of duration equal to the acceptable separation of signals. The trigger to the multivibrator passes through a diode in order to isolate the multivibrator when it is triggered, so that the second pulse will not interfere with the production of the trailing edge of the multivibrator pulse. The trailing edge is differentiated and shaped to look like the shaped signal; the two are then applied to the coincidence tube, which produces an output trigger when the second signal coincides with the delayed trigger from the multivibrator. As soon as the multivibrator recovers, the circuit is ready to accept another pair of pulses. It is thus important to keep the recovery time small. If the correct spacing of pulses is fairly long, the multivibrator may waste much time on single signals. Since the maximum triggering rate is the reciprocal of the code interval, many correctly coded signals may go unanswered. Operation is improved if several multivibrators are used in cascade to provide the code interval. Each one contributes only a fraction of the total delay, with the result that the maximum triggering rate is the reciprocal of the delay contributed by each multivibrator. Care should be taken to use multivibrators for which the delay does not

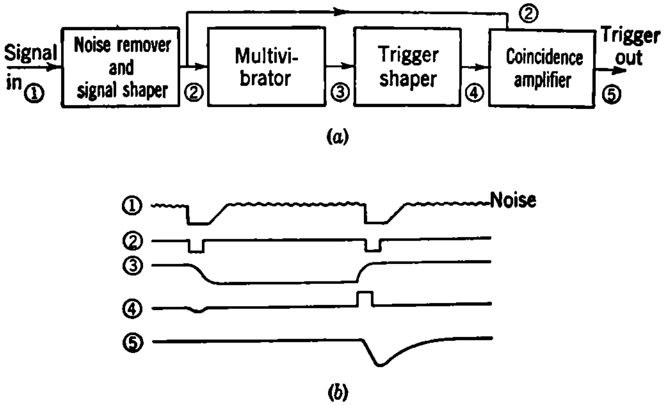


FIG. 9·14.—Double-pulse decoder using multivibrator delay. (*a*) Block diagram. (*b*) Waveforms at numbered points.

vary with repetition rate. A decoder in which multivibrators are used in cascade is considerably more complicated than a delay-line decoder.

**9·9. Multiple-pulse Decoders.**[1]—The methods of double-pulse decoding can be extended to apply to any number of pulses. The remarks made in Sec. 9·8 regarding the degrees of security of various methods of decoding apply equally well in this situation. Only the method described in detail in Sec. 9·8—accepting pulses the leading edges of which are separated by the correct intervals—will be treated here. The basic method, when used for multiple-pulse decoding, may be divided into two slightly different methods, according to whether delay lines or multivibrators are used.

*Two-by-two Coincidence.*—The first might be called the method of two-by-two coincidence. In this method, the first pulse is delayed by the correct amount for the first code space; if the second pulse coincides with the delayed first pulse, the second pulse is passed and delayed by the correct amount for the second code space, and so on. Each coincidence is only a double coincidence. A skeleton diagram of this type of circuit is shown in Fig. 9·15.

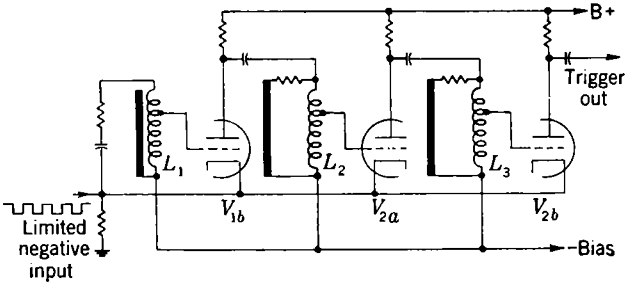


FIG. 9·15.—Four-pulse decoder, using two-by-two coincidence. $V_{1b}$, $V_{2a}$, and $V_{2b}$ are coincidence stages for the first, second, and third pair of pulses, respectively. The delay lines $L_1$, $L_2$, and $L_3$ determine the acceptable code intervals.

The signal at the input is negative and limited. It consists of four shaped pulses, of width $D$ μsec, with noise suppressed, coming from a low-impedance source. The signals are fed directly to the cathodes of all the coincidence tubes; this in itself, however, is not enough to drive the tubes to conduction. The first pulse is also applied to the input of the line $L_1$, which is matched at its input end, short-circuited at its output end, and has a tap at $D$ μsec from the input end. The negative first pulse travels down the line and is reflected as a positive pulse, gets back to the tap at the instant the second pulse is supposed to occur, and passes on to the input end of the line where it is absorbed. If the second pulse does occur when it is supposed to, coincidence occurs in the first tube and a negative

---

[1] Secs. 9·9 and 9·10 by C. L. Longmire.

pulse coincident with the second pulse of the signal is started into the second line $L_2$. The action is now repeated in this line; if the third pulse occurs at the right time, coincidence is achieved in the second tube, and so on. If all pulses have the right spacing, the last coincidence tube gives out a trigger. If, on the other hand, coincidence does not occur at any stage, the remaining stages do not act at all. The short section of line $D$ is necessary to prevent the reflected pulse from being canceled by the pulse with which it is supposed to coincide.

Codes may be changed by using plug-in delay lines or by using switches or relays to short-circuit the lines at various points. A large number of codes is possible: if there are three code spaces, and if each may have, for instance, any of five values, the total number of possible codes is $5^3$ or 125.

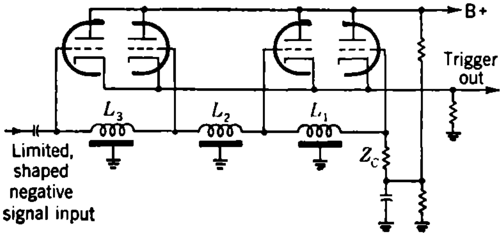


FIG. 9·16.—Four-pulse decoder, multiple coincidence.

If the code spacings are so large as to make electronic delay lines impracticable, multivibrators or supersonic delay lines may be used to obtain the delays.

*Multiple Coincidence.*—The second method of multiple-pulse discrimination might be called the method of multiple coincidence. In this method, each pulse is delayed long enough to make it coincide with the last pulse. A multiple-coincidence circuit is then used to produce a trigger only if all the delayed pulses coincide with the last pulse.

A skeleton diagram of this type of circuit is shown in Fig. 9·16. The negative input pulses, which must be limited and shaped, pass down the line and are absorbed in the resistance at the output end. $L_1$, $L_2$, and $L_3$ have delays equal to the proper first, second, and third pulse spacings respectively. If the signal has the right pulse spacings, then all of the grids will be driven negative together at the same time, and an output trigger will result. If all of the grids are not driven negative at the same time, however, there will be no appreciable output trigger. In this circuit the attenuation in the delay lines is cumulative; amplifiers may be needed, therefore, if the total delay required is fairly large. This method is best suited for short intervals.

The delay of both types of multiple-pulse circuits described in this section is such that the output trigger is timed with the last pulse of the interrogating signal.

*Coincidence Decoding.*—Coincidence decoding is a form of multiple-pulse decoding in which all pulses arrive at the same instant. The pulses must be received separately on different receivers. When received, they are applied to a coincidence circuit. It is usually necessary to suppress noise and to limit and shape the signals before they reach the coincidence tube.[1]

**9·10. Pulse-repetition-frequency Discriminators.**—It may be desired to accept only those pulses that arrive at the beacon with a certain repetition frequency. An obvious method for doing this is to delay each pulse by an amount equal to the correct repetition interval and to accept it if it coincides with the next pulse. Repetition intervals are usually so long as to prohibit the use of electronic delay lines. Other delay systems, therefore, must be used. Supersonic delay lines are admirably suited to this purpose.

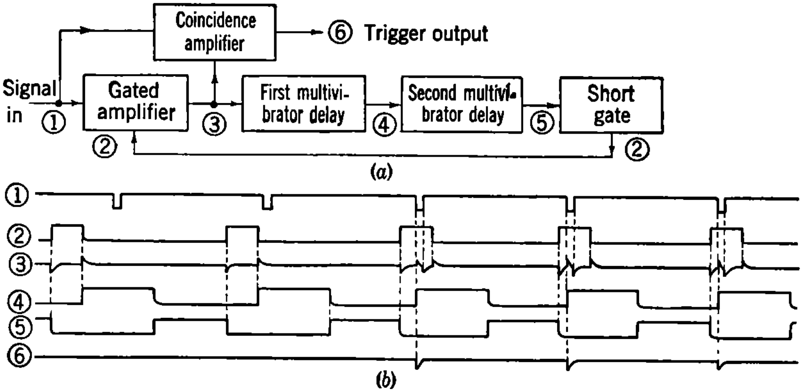


FIG. 9·17.—Simple pulse-repetition-frequency discriminator for beacons receiving only one interrogating signal. (*a*) Block diagram, and (*b*) waveforms at numbered points, showing how the discriminator locks on a signal of the correct pulse-repetition frequency.

Another method suitable for beacons designed to receive only one interrogating signal involves the use of multivibrators. Since the delay circuit must work for each pulse of the desired signal, and the required delay is equal to the time between signals, at least two multivibrators in cascade must be used to obtain the total delay and allow each one time

---

[1] For a thorough discussion of coincidence circuits refer to *Waveforms*, Vol. 19, Chap. 10, Radiation Laboratory Series.

to recover. Figure 9·17 shows a block diagram of a system using two multivibrators to select one set of received signals.

Suppose that a signal passed by the gated amplifier triggers the first delay multivibrator. When approximately half the desired repetition interval has elapsed, the first multivibrator triggers the second. This, in turn, triggers the short gate immediately before the next signal should occur. The short gate turns on the amplifier. If the next signal does occur at the proper time, it turns the amplifier off momentarily, thus triggering the first multivibrator again, and the cycle repeats. A coincidence circuit accepts a signal if it succeeds in passing through the gated amplifier.

If the signal does not occur within the short gate, the end of the gate turns the amplifier off again. This action triggers the first multivibrator again, and the cycle repeats itself. Obviously, if signals do not fall within the short gate, the circuit runs freely at a frequency slightly lower than that of the desired interrogating signal. Therefore, if a signal with the correct repetition frequency is present, the phasing of the signal with respect to the short gate will change continuously until the signal falls within the gate. If the difference between the free-running frequency and a triggering frequency is, for example, 10 cps, then the circuit should lock on a signal within $\frac{1}{10}$ sec.

To prevent locking-in on signals with repetition frequencies other than those desired, the length of the short gate is made such that two successive pulses with an incorrect repetition frequency would not fall within it. Because the desired pulse usually comes very close to the beginning of the short gate, the probability that an undesired signal will cut in even once is small. If a system including several of these circuits, each receiving a different pulse-repetition frequency, is planned, it is well to use a set of pulse-repetition frequencies no two of which have low common multiples.

This type of interrogation coding is not suited to scanning narrow-beam systems because of the long time necessary for locking-in. If the short gate is widened to increase the "searching" rate, the band of pulse-repetition frequencies that can lock in increases; consequently distinguishable frequencies must be farther apart.

Instead of two multivibrators to provide delay, several multivibrators may be used in cascade, each contributing a small amount to the total delay. If the final output trigger coincides with the next signal, an output trigger is obtained. Such a circuit has the advantage of being able to work on several recurring signals at once. If many interfering signals are present, however, they steal responses from signals with the correct repetition frequency; greater per cent response can then be obtained by increasing the number of multivibrators, and thereby decreasing the dead time. This circuit requires less time for a signal to break in than does that described above, except when interference is heavy. The delay multivibrator should be designed to have a fast recovery, so that the delay will be independent of the working rate. The complication required is such that the alternative method, which is the logical result of carrying this procedure to the limit, should be used in most cases; we refer, of course, to the use of a storage method, using supersonic delay lines.

# CHAPTER 10

# RESPONSE CODERS

By C. L. Longmire

In this chapter we consider coders used for identifying beacons. Codes and coders for other purposes, such as communication, data transmission, and remote control, are discussed in Chap. 11.

**10·1. General Considerations.**—One of the most useful attributes of a beacon is that it can identify itself. In air navigation, the operator of an airborne interrogator must know the geographical location of a beacon so that the data he obtains from it will have a reference point; the beacon must therefore locate the reference point by identifying itself. Conversely, if a ground interrogator is controlling one of several beacon-carrying aircraft, the operator must be able to single out on his scope the particular aircraft he desires to control; the airborne beacon in which he is interested must be able to make itself distinguishable from beacons carried by other aircraft in the vicinity. In short, it is important that beacons have "names" just as people have names.

A beacon can identify itself only by its response. A beacon that does not respond is of no more value in identifying its particular location than "no beacon" would be; codes that interrupt the response of the beacon are, therefore, undesirable. The mere presence of a response is highly identifying, regardless of its form, because it indicates that the beacon was able to respond to the particular form of interrogating signal used.

A beacon is partially identified by coded interrogation, which was discussed in the preceding chapter from a different standpoint. The beacon decoder was there regarded as a circuit to establish the identity of the interrogator from which signals come, instead of as a means of identifying the beacon. Interrogation coding as a means of identification has the disadvantage that many—all but one, usually—of the reply codes consist of no response.

In response coding, just as in interrogation coding, significance can be attached to several characteristics of the signal. The radio frequency of the response almost always identifies the beacon to some extent. The principle involved is obvious so that nothing further need be said of this type of coding.

Codes based on variation of pulse amplitude usually are not useful because they involve a reduction in the range of the system. Furthermore, variations in the strength of various signals received by the interrogator would necessitate a fairly complex automatic-gain-control system to prevent limiting of the beacon signal in the interrogator receiver. A third disadvantage of amplitude coding is that the best source of high-power microwaves—the magnetron—is not at present capable of reliable amplitude modulation.

The repetition rate of the beacon response cannot be varied in the same way as that of an interrogator because the beacon depends upon the interrogator for its repetition rate. By causing the beacon not to respond to all signals, however, the response rate can be made to equal submultiples of the interrogation rate. For instance, a particular beacon might be identified by the property of answering only every third interrogation from a given interrogator. It would be easy to make a beacon respond in this way if only one interrogator were triggering it, but it would be more difficult if several interrogators were triggering it at the same time.

The pulse width of the reply is usually available for coding. Because it is desirable to keep the pulse width short to save power, it becomes necessary to use fine gradations of pulse width in order to obtain an appreciable number of codes. An expanded sweep is then necessary for interpreting the code visually. For easy visual interpretation, pulse-width response coding is less desirable than multiple-pulse coding. If the decoding or interpretation is done by circuits, however, this is no longer true. For pulse-width coding, the interrogator-receiver must be designed carefully to avoid stretching the signal.

A multiple-pulse code has three identifying features: number of pulses, spacings of pulses, and pulse widths. Codes of varying complexity can be constructed by using these three features. The simplest type is one in which number of pulses only is significant, and all spacings and pulse widths are the same. The most complex is one in which the number of pulses, each code space, and each pulse width are assigned one of a large set of values. Use of this method yields many identifiable codes. (See Sec. 5·17.) A system of medium complexity that has found considerable use allows each space to be either "long" (35 μsec) or "short"(15 μsec), varies the number of pulses from two to six, and makes all pulses the same width (0.5 μsec).

Fine gradations of spacing and pulse width can be decoded better by circuits than by visual interpretation because a code that is to be interpreted visually must have features that are resolvable on the interrogator scope. The spacings given in the preceding paragraph were chosen to make a code just resolvable on a 100-mile sweep on a 5-in. PPI. From the size of the required spacings, it is plain why significance was attached to spacing but not to pulse width. If the interrogator has an expanded sweep that can be started at any range, however, shorter codes involving pulse width are feasible for visual interpretation.

**10·2. Single-pulse Response Coders.** *Pulse-width Coders.*—In all pulse modulators, a circuit element fixes the duration of the transmitted pulse,[1] whether pulse-width coding is to be used or not. Thus, all modulators contain some of the essential elements for pulse-width coding.

The simplest type of pulse-width coding is that in which the beacon responds with one of a given group of pulse widths. As it may be desirable to change the code (or pulse width) of a beacon from time to time, the circuit that fixes the pulse width should be equipped with a multiposition switch that has a position for each pulse width. In line-controlled blocking oscillators and thyratron pulse-forming circuits, the multiposition switch controls the length of the delay line. Multivibrators, which are often used to fix the pulse width when fairly long pulses (greater than 2 or 3 μsec) are used, are easily controlled by varying a voltage or a resistance. In all cases, the tolerance of the various pulse widths must be less than the minimum difference in pulse widths for different codes.

Another method of pulse-width coding is that of varying the pulse width in time according to a definite plan. For visual interpretation, the pulse width may be alternated between long and short values, the ratio of these values being 2 or 3 to 1. The pulse width may be "modulated" automatically in the desired manner by a motor-driven cam or coding wheel. Obviously, the decoding time in such a system is a matter of several seconds and even longer if there are many complex codes. This method is, therefore, not suited to scanning interrogators. Morse-code characters are universally used in practice.

Because of the long time necessary to transmit all the elements of a code, pulse-width coding is often called "slow" coding.

It is possible, by obvious means, to present the code visually, in the form of a blinking light, or aurally, as ordinary radio code signals.

## MULTIPLE-RESPONSE PULSE CODERS

A multiple-pulse coder usually consists of three parts: a blanking gate, trigger delay stages, and a collecting and amplifying circuit.[2]

The trigger delay stages actually form the code, one pulse at a time; the last step is to collect and combine the individual pulses and perform the necessary shaping. The collecting methods depend somewhat upon the method of obtaining the delay; they vary for different types of delay stage and will be discussed in connection with each.

---

[1] See Chap. 12.

[2] For a complete discussion of the circuits used for generating the waveforms used in response codes, see *Waveforms*, Vol. 19, Radiation Laboratory Series.

**10·3. Blanking Gate.**—The blanking gate is used to prevent extraneous signals from interfering while the coder is forming a code. The term "blanking gate" has been used to denote both the circuit and the waveform that it produces. Sometimes the last stage of the component preceding the coder may act as the blanking-gate generator; in some instances of low working rate a blanking gate may not be necessary, because of the small probability of interference. Duty-ratio limitation is often secured by varying the duration of the blanking gate.[1]

The blanking-gate generator usually consists of a multivibrator that produces a rectangular waveform longer than the code, and has as short a recovery time as is possible practically; therefore the probability of occurrence of abnormal gate lengths is small.[2] The multivibrator is put into the circuit so that the incoming trigger sets it off, and the leading edge of its waveform is then used in turn to trigger the first delay stage. Care must be taken in the design to avoid the possibility of firing the first delay stage on the input trigger even when the multivibrator does not fire; this might happen, for example, if the multivibrator has been triggered just previously.

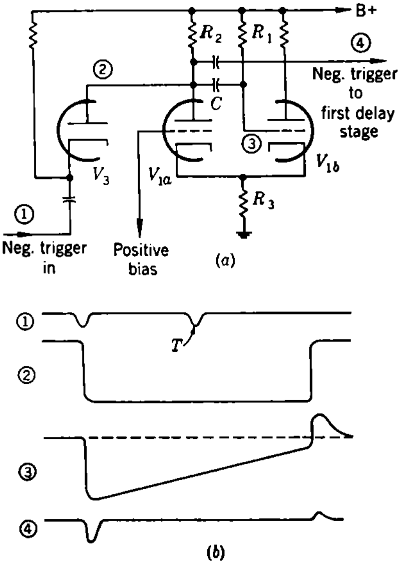


FIG. 10·1.—(a) Blanking gate generator with diode input. (b) Waveforms at numbered points of (a). T—rejected trigger pulse.

One simple method that guarantees correct action is to pass the input trigger through a diode that is biased off as soon as the gate fires. (See Fig. 10·1.) In this circuit the triode $V_2$ is normally on and $V_1$ normally off, and the voltage across the diode $V_3$ is practically zero. The first incoming negative trigger is passed by the diode and triggers the multivibrator. The plate of the diode is pulled far negative with respect to the cathode during the rectangular multivibrator waveform, so that additional input triggers do not get through the diode. If it were not for the diode, additional triggers would be passed on to the first delay stage. The product $R_1C$ determines the gate length, other parameters being

---

[1] See Sec. 6·4.

[2] Abnormal gate lengths occur if a multivibrator is triggered before it has completely recovered.

fixed. The product $(R_2 + R_3)C$ determines the recovery time, independent of $R_1$. It is therefore well to use a large value of $R_1$ and a small value of $C$ to obtain both the proper gate length and quick recovery.

**10·4. Coders for Equally Spaced Pulses.**—Two methods have been used to generate a given number of equally spaced pulses for each input trigger. A skeleton diagram of one method is shown in Fig. 10·2. In this circuit the blocking oscillator $V_2$ is normally biased off. The incoming trigger is amplified by $V_1$ and triggers the blocking oscillator $V_2$, which regenerates and produces a pulse the length of which is determined by the delay line $L_1$ in the grid circuit. A negative pulse is fed into the delay line $L_2$ connected to the plate of the blocking-oscillator tube. The negative pulse travels down the line, is reflected with a reversal of sign at the short circuit, and travels back to the plate where the impedance presented is essentially that of an open circuit. The pulse is therefore reflected without a change in sign. It then makes another trip down the line and back, coming back as a negative pulse this time. This negative pulse

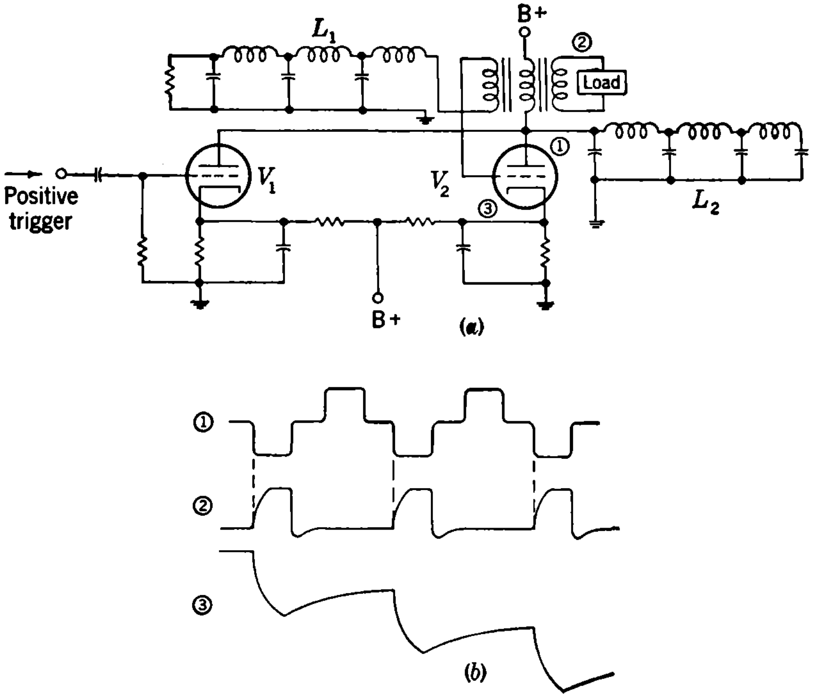


FIG. 10·2.—(a) Blocking oscillator for generating equally spaced pulses. $L_1$—delay line for controlling pulse width. $L_2$—delay line for controlling pulse spacing. The delay in $L_2$ is one-fourth the pulse spacing. (b) Waveforms at numbered points.

triggers the blocking oscillator again, and the action is repeated. After a time that depends on the size of the cathode condenser, the cathode voltage rises to a point at which the doubly reflected pulse is not large enough to initiate regeneration. The reflections then die out in the plate line, and the cathode voltage returns exponentially to its normal value. Meanwhile, the proper number of properly shaped pulses have been delivered to the load, which may be a power amplifier to drive the transmitter.

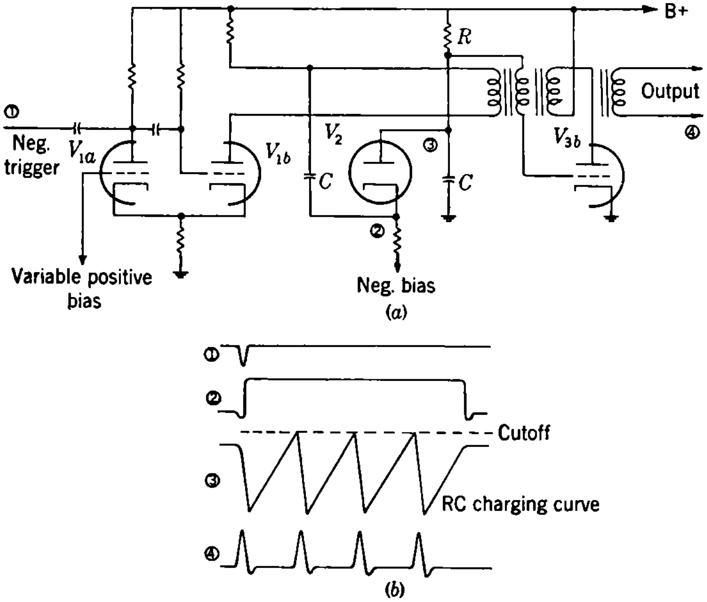


FIG. 10·3.—(a) Gated blocking oscillator for generating equally spaced pulses. $V_{1a}$, $V_{1b}$—multivibrator; $V_2$—diode; $V_{3b}$—blocking oscillator. (b) Waveforms at numbered points in (a).

The slow return of the cathode voltage is detrimental in that closely spaced additional input triggers are apt to get responses that contain fewer pulses than the correct code. A less economical coding method that does not have this disadvantage is to generate a gate of the proper length and to allow a blocking oscillator to run freely during the gate. A circuit of this type is shown in Fig. 10·3.

In this circuit, the incoming trigger fires a multivibrator $V_1$, $V_2$, which produces a positive gate the length of which is controlled by the variable positive bias. The diode $V_3$ normally conducts and so causes the blocking oscillator $V_4$ to be cut off. The leading edge of the gate triggers the blocking oscillator, which regenerates with the result that the grid is left with a large negative voltage. This voltage rises exponentially toward the plate-supply voltage with the time-constant $RC$; it now passes its normal level because the diode cathode voltage has been raised by the gate. Thus the grid enters the conducting region and the blocking oscillator fires again. This action is repeated until the gate ends and no longer allows the blocking-oscillator grid to enter the conducting region. The number of pips can be changed by changing the gate length. The multivibrator should have a short recovery time.

It is evident that the above circuits do not require pip-collecting circuits, although shaping and power amplification may sometimes be required.

## PULSE-SPACING CODERS

**10·5. Delay-line Delay Stages.**—When the spacings are small, less than about 10 μsec, delay lines can be used to obtain the code. Figure 10·4 shows a skeleton circuit. Each of the tubes is normally biased to cutoff. The incoming positive trigger turns on the first amplifier, which starts a negative pulse into the first delay line; this pulse is reflected and reappears as a positive pulse at the input terminals, where it is absorbed. The reflected positive pulse turns on the second amplifier, and so on. A common resistor in the cathode circuits of all the tubes collects the pips forming the code. These pips may be amplified and shaped if necessary.

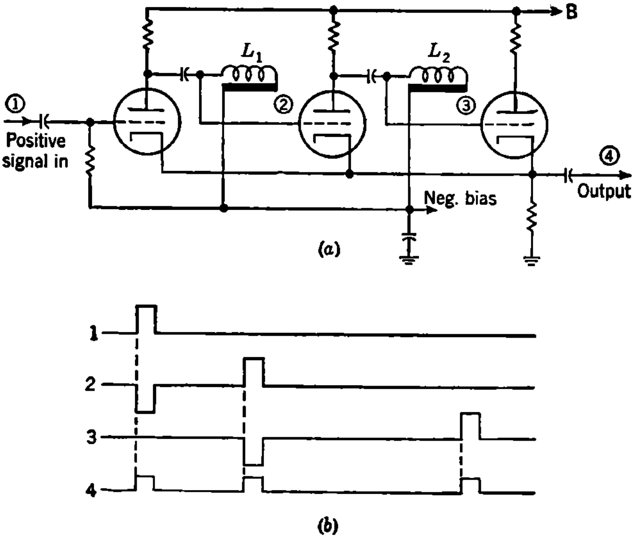


FIG. 10·4.—(a) Three-pulse delay-line coder. Delay of $L_1$ = half of first code space. Delay of $L_2$ = half of second code space. (b) Waveforms at numbered points of (a).

Each code space can be varied by changing the position of the short circuit in the delay line corresponding to that space.

The circuit described above has two advantages: it uses each line to get twice the normal delay, and the amplifiers counteract the effects of attenuation in the lines. These advantages over the simplest type of circuit, consisting of one long line with taps to take off the pulse with various delays, are well worth the additional complications.

**10·6. Multivibrator Delay Stages.**—Cascaded biased multivibrators may be used economically to obtain the delays necessary for pip coding for a wide range of code spaces—from 2 or 3 $\mu$sec up. In this type of circuit, each multivibrator except the first is triggered by the trailing edge of the rectangular wave produced by the preceding multivibrator. The first multivibrator is triggered by the leading edge of the blanking gate. Small triggers are produced by differentiating the trailing edge of the rectangular wave produced by each multivibrator, and the leading edge of the first rectangular wave; these are collected, amplified, and shaped.

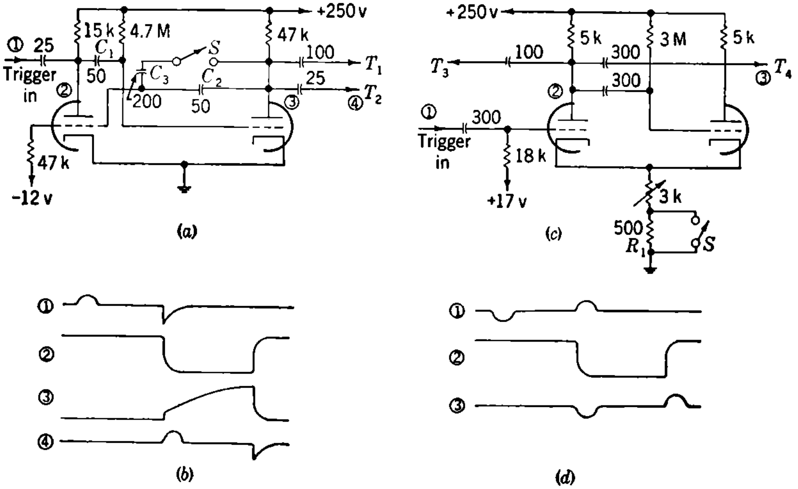


FIG. 10·5.—Two types of multivibrators used in pip coders. (a) Grid-plate-coupled type. $T_1$—trigger to collecting line; $T_2$—trigger to next stage. (b) Waveforms at numbered points in (a). (c) Cathode-coupled type. $T_3$—trigger to collecting line; $T_4$—trigger to next stage. (d) Waveforms at numbered points in (c). NOTE: In both circuits capacitances are in $\mu\mu$f. Values are such that the delay with $S$ open is 12 $\mu$sec; with $S$ closed it is 40 $\mu$sec.

Two general types of multivibrator have been used: the plate-grid-coupled type and the cathode-coupled type. Both are shown in Fig. 10·5. In the plate-grid-coupled type the delay is changed by switching in additional capacitance between the normally on plate and the normally off grid. The circuit shown gives a spacing of about $\frac{1}{2}$ μsec with $C_3$ out and 40 μsec with $C_3$ in. The sizes of $C_1$ and $C_2$ are such that $C_2$ controls the length of the short space, and $C_1$ controls the length of the long space; a trimmer capacitor across each of these makes long and short spaces adjustable separately.

In multivibrators of the cathode-coupled type, the spacing is increased by switching out a portion of the cathode resistance. Errors in any circuit component except $R_1$ have proportionate effects on the short and long spaces so that adjustment of the potentiometer tends to correct both short and long spaces, although not equally well.

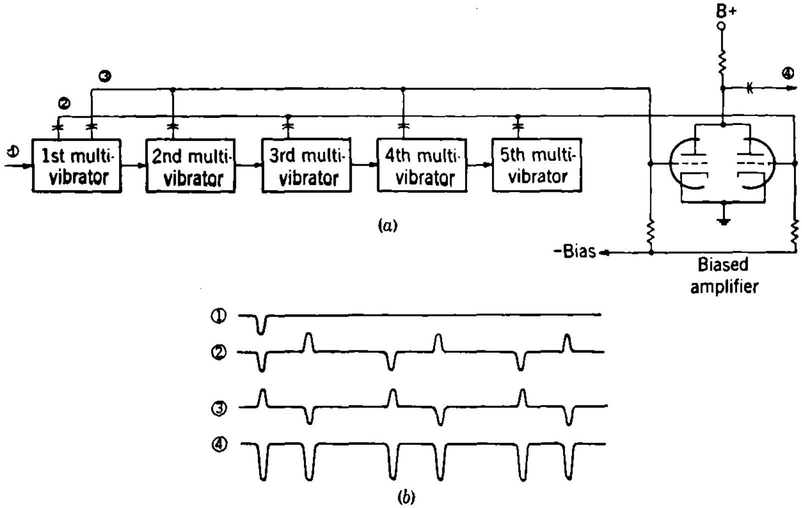


FIG. 10·6.—Collecting network for cathode-coupled multivibrator coders. (a) Block diagram, showing collecting stage. (b) Waveforms at numbered points in (a).

For the particular circuits shown, multivibrators of the plate-grid-coupled type require much less plate current than those of the cathode-coupled type, although this need not be a general rule.

Figure 10·6 shows how the pips may be collected in a circuit using the cathode-coupled multivibrators of Fig. 10·5. The condensers couple to the first or second plate (in Fig. 10·6), according to their position relative to the block representing each multivibrator. Two collecting lines are necessary to avoid canceling of simultaneous positive and negative triggers. The trigger on one line is positive when the trigger on the other line is negative; as a result, either one or the other of the biased amplifiers is driven to conduction. The output pulses of the double amplifier are usually amplified, inverted, and passed through a cathode follower to give a low output impedance.

In collecting pips from a multivibrator of the plate-grid-coupled type, advantage may be taken of the almost triangular waveform at the second plate. On differentiation, this waveform gives only a small positive trigger at the leading edge and a considerably larger negative trigger at the trailing edge; this allows collecting of all triggers on one line, as is shown in Fig. 10·7, because the negative triggers override the positive ones.

In the multivibrator circuits shown, the spacing is controlled by means of switches. Continuous electronic control of the spacing may be desired in some instances. For this purpose, control of the voltage of the normally-off grid in the cathode-coupled multivibrator gives approximately linear control of the spacing.

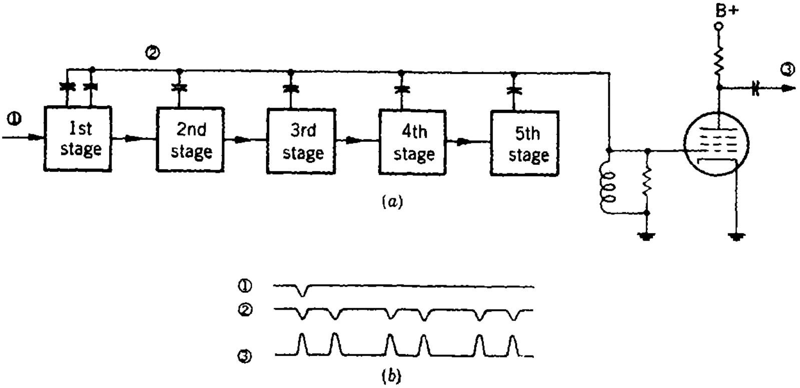


FIG. 10·7.—Collecting network for grid-plate-coupled multivibrator coder. (a) Block diagram. (b) Waveforms at numbered points in (a). Note the use of a single collecting line.

The number of pips can be reduced simply by biasing off or removing plate voltage or ground from the later delay stages.

**10·7. Overshoot Delay Stages.**—A delay circuit of the overshoot type uses a single triode section to obtain each delay and is based on the formation of overshoots by differentiation. Figure 10·8 shows a skeleton diagram of two stages.

In this circuit the tubes are normally at zero bias, and the resistor $R_1$ is much larger than the zero-bias resistance of the tube; consequently very little voltage drop appears across the tube. Assume that the current in $V_1$ is shut off for a period $T_1$. The waveform at the plate of $V_1$ is then as shown, the time constant of the rise of the pulse being $R_1C$. The grid resistance of $V_2$ is much smaller than $R_1$ and is neglected. Since $R_1C$ is somewhat less than the shortest desired delay, the maximum amplitude of this pulse is relatively independent of $T_1$. During $T_1$ the grid voltage of $V_2$ does not rise much because of the low grid resistance.

But at the end of $T_1$, the grid of $V_2$ is swung negative far beyond cutoff. However, the grid voltage immediately starts to rise exponentially toward the plate-supply voltage at a rate determined mainly by $C$, $R_2$, $R_3$, and the zero-bias resistance of $V_1$ but stops and remains at zero. Thus, plate current in $V_2$ is turned off for a period $T_2$ immediately following $T_1$. The magnitude of $T_2$ is independent of $T_1$ but can be changed by short-circuiting $R_3$. At the plate of $V_2$, another positive rectangular pulse is developed, which can be used for driving the next stage.

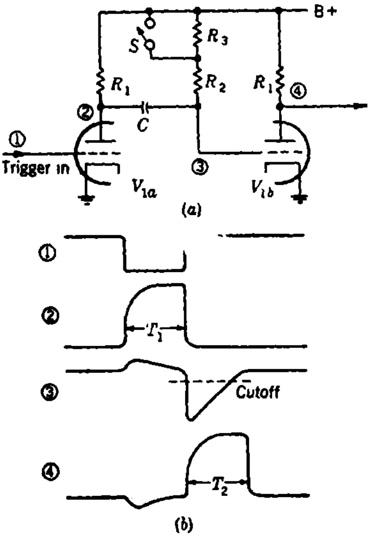


FIG. 10·8.—(a) Two stages of an overshoot delay circuit. (b) Waveforms at numbered points of (a). Note that the input trigger controls the first space. S changes the code space value.

The amount of delay per stage can also be varied by changing the voltage to which the grid resistance is returned, instead of changing the resistance itself. This method is advantageous when the code spaces must be controlled from a distance through cables; it is not subject to the distortion of a high-impedance waveform by the capacitance of the cables.

It is evident that in a circuit of this type the first delay stage cannot be triggered in the usual manner. A negative rectangular trigger pulse is required. Such a waveform can be obtained from the blanking gate which usually precedes the delay stages.

Tubes with low zero-bias resistance and with high $g_m$ are required; otherwise the waveform degenerates from stage to stage. This type of circuit is better when the number of stages in succession is small, because trouble results from the small positive part of waveform 3 when the amplification is large. Three stages of 7F8 tubes work well to give a four-pip code.

*Pip Collection.*—An attempt to collect in the same manner as for multivibrators will fail because the extra loading is injurious to the waveform at each stage. However, the method shown in Fig. 10·9 has been used successfully. Here the first and third positive rectangular waves are applied to the grids of a double amplifier through the networks consisting of $R_1$, $R_2$, and $C$. The high resistors $R_1$ and $R_2$ do not load the plates of the delay stages; $C$ is small and is added to prevent degeneration of the waveform. Across the primary of the transformer $T_1$ there is a trigger, positive or negative, at the beginning and at the end of each delay. Two secondaries of $T_1$ are connected so that one or the other of the blocking-oscillator grids is brought into the conducting region regardless of the sign of the trigger across the primary of $T_1$. When regeneration begins in one tube of the blocking oscillator, it soon spreads to the other tube, so that both tubes contribute to the pulse. The recovery time of the bias network for the blocking oscillator is less than the minimum spacing of the code pips.

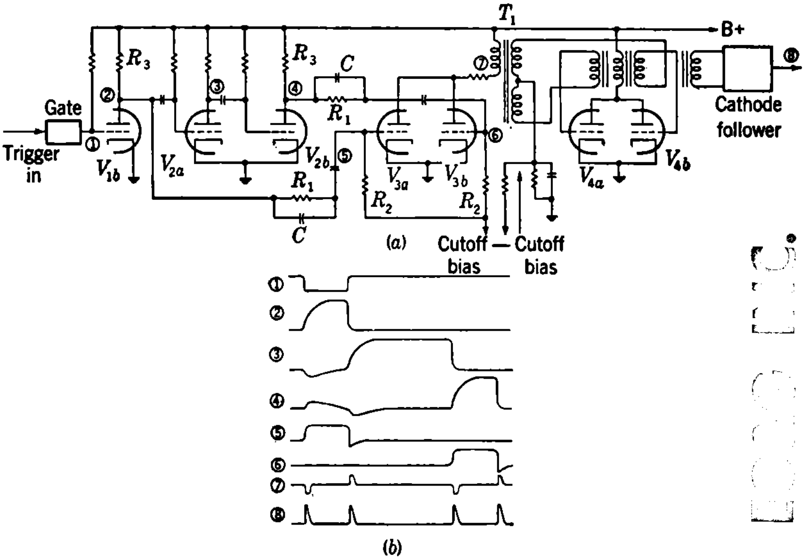


FIG. 10·9.—(a) Pip collecting in overshoot coder. $V_{1b}$, $V_{2a}$, $V_{2b}$—delay stages; $V_{3a}$, $V_{3b}$—collecting stages; $V_{4a}$, $V_{4b}$—blocking oscillator. (b) Waveforms at numbered points of (a). NOTE: All tubes are 7F8. With $R_1$ = 470K, $R_2$ = 270K, $R_3$ = 56K, $C$ = 47 $\mu\mu$f, short spaces are 12 $\mu$sec.

The overshoot type of coder is a little more economical than the multivibrator type. There are, however, as mentioned above, limitations on the possible number of code pips, and particular tube types are required.

**10·8. Blocking-oscillator Delay Stages.**—Blocking oscillators can be used economically to form the basis of delay stages. (See Fig. 10·10.) In this circuit a positive trigger is required for the first stage. The constants of the grid circuit of $V_1$ should be chosen with due consideration for the circuit supplying the initial trigger. The blocking oscillator $V_1$ is normally biased off. When fired by the initial trigger, it draws a short, heavy current pulse that charges $C_3$ in the plate circuit. The current pulse flowing through $R_4$ produces the first pip.

As soon as the pulse is over, the resonant circuit composed of $C_3$ and the 85-mh choke starts to oscillate. The value of $R_2$ is chosen to give the correct amount of damping. The second biased-off blocking oscillator $V_2$ fires on the positive rise of this waveform. The grid circuit of $V_2$ does not recover fast enough to fire twice.

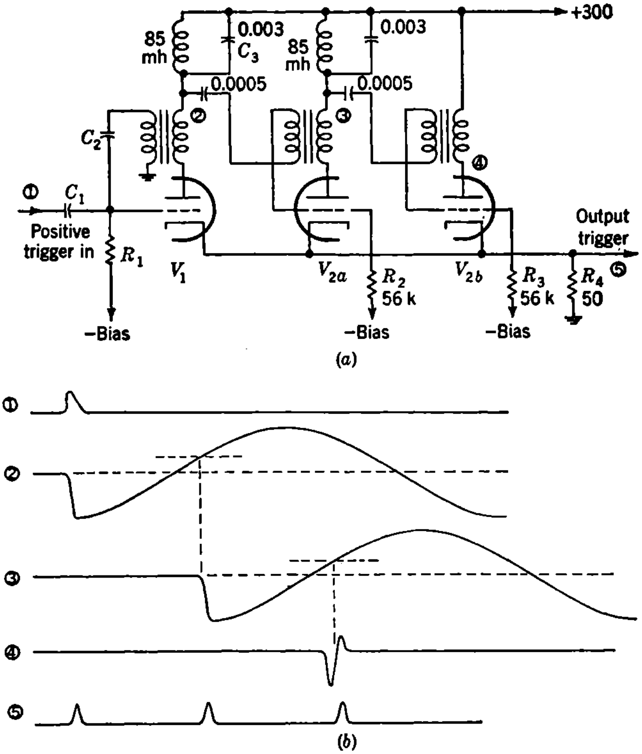


FIG. 10·10.—Three-pip blocking oscillator coder. (a) Circuit diagram. Tubes are all 7F8. (b) Waveforms at numbered points in (a). Note that no special collecting circuit is required, average current is low, a minimum of tubes is used, and the output voltage is large and from a low impedance source.

The resonant circuit in the plate of $V_2$ then oscillates in the same manner as the first stage and fires $V_3$, which does not require a resonant circuit because it produces the last pip.

The spacing between the positive output pips may easily be changed by switching the values of the condensers in the plate tank circuit and by adjusting the bias of the blocking-oscillator grids. With the constants shown, the bias values may be adjusted to give spacing of about 45 μsec.

The circuit has several advantages. All tubes are normally nonconducting and the average plate current drawn is very low even for high working rates—less than 2 ma for 4 pips at 2000 responses per second. Only one triode is required per pip, and the collecting requires no special circuit. The output consists of sharp triggers from a low-impedance source.

## OTHER TYPES OF CODER

**10·9. Miscellaneous Pip Coders.** *CRT-Pip Generators.*—An interesting method has been proposed for generating pip codes by means of a special tube. The method has not been used in beacons but has been used for similar purposes in other pulse equipment. The beam of a cathode-ray tube is deflected by a linear sweep circuit that is triggered by the input signal. The tube face has a number of evenly spaced electrodes so placed that the beam is swept across them.[1] The electrodes are grounded through a common resistor, across which there appears a trigger each time the beam strikes an electrode. These triggers may be amplified and shaped to form a pip code. The spacings may be changed by changing the sweep voltage and by selection of appropriate electrodes.

*Linear Sweep Delays.*—A method similar to the one just described, but which has not been thoroughly developed, is one in which one linear voltage rise is used as the source of all the delays. The voltage sweep is triggered by the input trigger, and pulses are generated at the instants when the voltage passes certain preset levels.

The voltage-level-selecting circuit is the main problem. Unless each pip can be obtained with one triode, or less, the circuit is not economical. An arrangement using neon bulbs could be used, each bulb breaking down at a different value of the sawtooth voltage; but it has been found that bulbs available at present do not break down at the same voltage on every sweep. Standard "pickoff" schemes involve at least a double triode.

*Low-power Coders.*—Low power drain is often of prime importance as, for example, in battery-operated beacons. Power is consumed mainly for two uses: filament supply and plate supply. Considerable power can be saved by choosing tubes with low filament-power requirements, although such tubes frequently do not have other desirable characteristics.

Some circuits require fewer tubes and less plate power than others; the blocking-oscillator type described above in Fig. 10·10 is best from this

---

[1] A general discussion of such devices is given in *Waveforms*, Vol. 19, Chap. 3, Radiation Laboratory Series.

standpoint. The tubes shown in Fig. 10·10 do not have low-drain filaments; low-drain tubes would have to be substituted to make a low-power coder. The tube type chosen should have a sharp and uniform grid-cutoff characteristic to insure stability of the blocking oscillators. In the past, no low-drain tubes with the necessary sharp cutoff have been available. Large variations in cutoff characteristics from tube to tube make it difficult to adjust the bias to a value such that no tubes oscillate spontaneously, and all tubes operate on the input trigger. Both 3A5 and 3B7 tubes have been used; neither type is completely satisfactory. Sacrifices must often be made in the total number of codes and in performance to get lower drain.

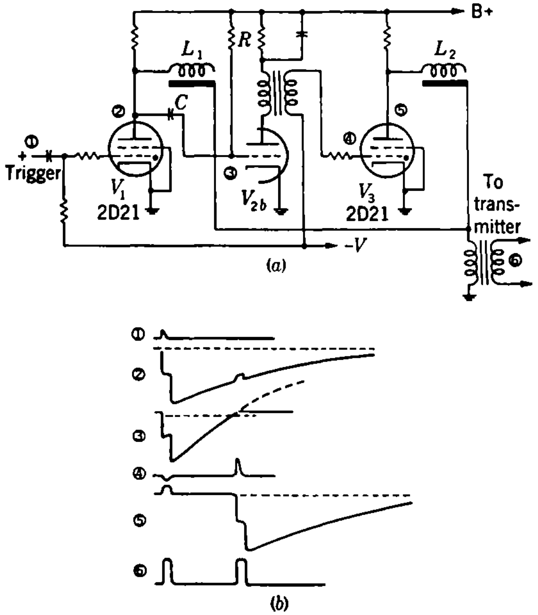


FIG. 10·11.—(a) Two-pulse combined coder-pulser for a low-power transmitter. $V_1$, $V_3$—pulser stages; $V_{2b}$—delay stage. $L_1$ and $L_2$ control the first and second pulse widths respectively. *RC* controls the spacing between pulses. (b) Waveforms at numbered points of (a).

**10·10. Coder-pulser Combinations.**—Usually the code is generated in more or less "rough" form in the coder and passed on to the pulser for power amplification and pulse shaping. Special circuits in which the tube generating the code delivers enough power with the proper waveform for modulation can be designed, however, at least for simple codes and low-power transmitters. Such circuits may permit a considerable saving of tubes, space, and power.

For pulse-width coding it is merely necessary to use the pulse-width determining circuit element with a tube that will supply the necessary power. This is, in fact, usually done, making coder and pulser indistinguishable. A familiar method is the use of a thyratron to discharge a delay line of variable length.[1]

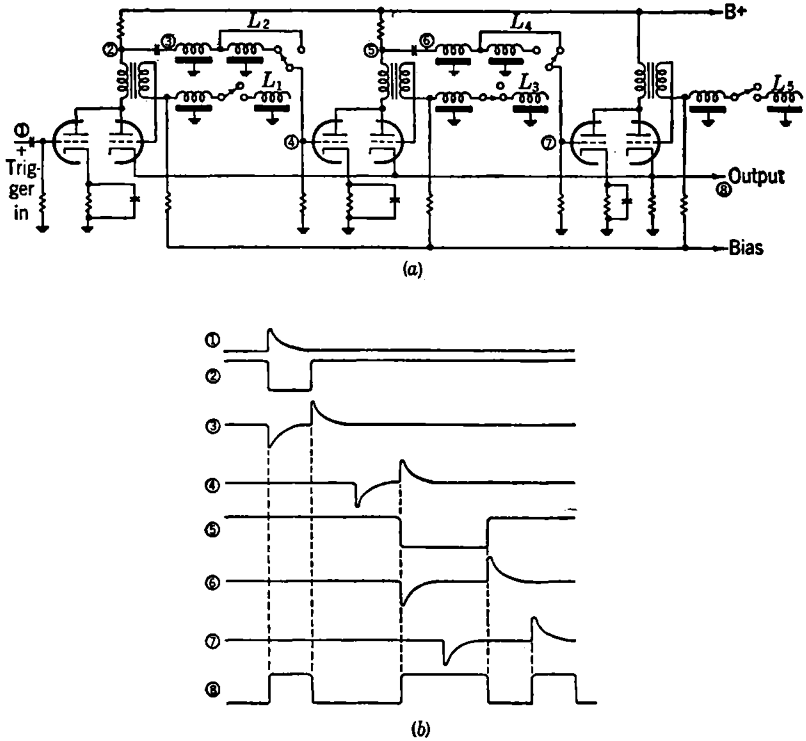


FIG. 10·12.—(a) Combined pulse-width and pulse-spacing coder. $L_1$—delay line that determines first pulse width; $L_2$—delay line that determines first space; $L_3$—delay line that determines second pulse width; $L_4$—delay line that determines second space; $L_5$—delay line that determines third pulse width. Switches control the delay-line constants. (b) Waveforms at numbered points in (a).

Figure 10·11 shows a combined coder-pulser for multiple-pulse codes, using 2D21 thyratrons to generate the pulses. The delay is obtained by differentiation of the "step" waveform resulting from discharge of the line associated with the first pulse. The delay lines determine the width of the pulse. More stages can be added to give more than two pulses.

**10·11. Combined Pulse-width and Pulse-spacing Coders.**—If the interrogator sweep is fast enough to resolve differences in pulse width,

---

[1] See Chap. 12.

multiple-pulse codes using both pulse spacing and pulse width can be used. For the same number of pulses, a code of this type has one more than twice as many variables as a pulse-spacing code, and the number of possible codes is greater than the square of the number of a pulse-spacing code. Figure 10·12 is an example of a circuit readily permitting control of both pulse width and pulse spacing.

In order to conserve power and minimize display space, pulse widths and pulse spacings will usually be short—a few microseconds at most; delay-line circuits will probably, therefore, be preferable to multivibrators. In Fig. 10·12 each pulse is generated by a line-controlled blocking oscillator. A rectangular voltage pulse occurs across a resistor in the plate circuit; this rectangular pulse is differentiated, giving a positive trigger at the end of the voltage pulse. The positive trigger is passed through a delay line of length adjusted to give the correct spacing, and then triggers the next stage. The output pulses are already collected, and come from a low-impedance source. In amplifying, care should be taken to preserve pulse width and spacing.

# CHAPTER 11

# BEACONS AS PULSE COMMUNICATION SYSTEMS

By G. O. Hall, C. L. Longmire, R. A. Minzner,
M. D. O'Day, and A. Roberts

**11·1. Beacon Communication Systems.**[1]—In previous chapters, beacons have been considered from the standpoint of their use as navigational devices. In this chapter, methods by which the interrogator-beacon system can be used as a pulse-communication system are discussed. The use of such methods is limited to systems in which beacon interrogation is continuous. No communication methods have as yet been devised for systems in which the interrogator scans and the beacon is interrogated intermittently.

Bandwidth requirements discussed in Sec. 5·18 must be met before voice-communication systems can be built. A pulse system has a bandwidth adequate for voice communication only if its pulse-repetition frequency is high or if each pulse contains more than one element of information.

It is legitimate to question the advantage of adding complications to a beacon system in order to provide an auxiliary communication system. There are reasons for doing so. For example, very often the saving in power and weight over another communication system makes the use of a beacon-communication system desirable. The weight added to the beacon may be only that of a code-sending key, or, if equipment making complex modulation possible is added, it may amount to 10 or 20 lb. Considerable security may be provided. A distinct advantage is that the interrogator operator can communicate with the operator of any particular beacon he is interrogating.

Another advantage of a beacon-communication system is the possibility of getting a high signal-to-noise ratio. The power margin available in a beacon system may be large compared with that available in a pulse-communication system, especially when radar sets are used as interrogators.

It is true, however, that the traffic capacity of a beacon may be reduced while it is being used as part of a communication system. Not only will it be unable to answer as many interrogations as it could other-

---

[1] By M. D. O'Day.

wise, but the communication replies may cause the type of clutter called "fruit." It is, therefore, not often desirable that a fixed long-range navigational ground beacon be used for communication that requires high repetition rates, because such a beacon may have to respond to the routine interrogations of a large number of aircraft.

Many applications in which beacons are used, however, do not utilize their maximum traffic capacity; this is often true in many uses of airborne beacons and portable ground beacons. A communication system may be of advantage in connection with such beacons in definitely identifying the beacon operator and permitting moderately secure communication between him and the interrogator operator. Even when security is not important, it can be advantageous to be sure that one is talking to an operator associated with a specific target on the interrogator scope.

## COMMUNICATION METHODS NOT INVOLVING SPEECH TRANSMISSION

BY C. L. LONGMIRE AND G. O. HALL

The purpose of a communication system is to transmit data that may vary widely in complexity. They may be simple, indicating only whether the state of a system is *A* or *B*, or they may be as complex as speech itself. At the moment, we are concerned with methods of transmitting data that require less bandwidth than is necessary for voice communication. Pulse-repetition frequencies lower than those that must be used in voice work are useful.

The functions of modulation and interpretation may be performed by operators, by circuits, or by combinations of operators and circuits. The treatment that follows is divided according to the methods used.

### METHODS OF MODULATION

Of the large number of possible methods of modulating a series of pulses, we will discuss only a few in detail. Pulse-amplitude modulation is undesirable if one wishes to obtain the best possible signal-to-noise ratio; it is also difficult to apply to magnetrons of current design. In the methods to be discussed the pulse power does not vary, although the average power may do so.

**11·2. Modulation of the Pulse-repetition Frequency.**—The pulse-repetition frequency is often one of the easiest parameters of a pulse transmission to change, and there are simple and effective methods available for modulating it. A reactance tube may be used to vary the frequency of an oscillator. More often, the length of the cycle of a multivibrator or other square-wave generator that controls the repetition period is varied by varying a grid voltage.

Pulse-repetition-frequency modulation is best adapted to hard-tube pulsers. It is not often feasible in interrogators using resonant charging of pulse-forming networks in the pulser.

**11·3. Space Modulation.**—A method of communication from interrogator to beacon is illustrated in Fig. 11·1. In this method, $2n$ pulses are sent out per second, but with a spacing somewhat as illustrated as in (*b*) of Fig. 11·1. The whole array of pulses may be considered as the superposition of two sets of regularly spaced pulses of $n$ per second each, the pulse interval being $T = 1/n$, but with the two sets displaced relatively by a time, in general, different from half of this interval. Each of the partial sets of pulses may be represented by a Fourier series having terms corresponding to frequencies of 0, $n$, $2n$, $3n$, etc. By use of a suitable filter the component of frequency $2n$ can be isolated. Each of the two sets of interlaced pulses will contribute a component oscillation of the same amplitude at this frequency, the resultant depending on the relative phases. A displacement in time of the two sets of pulses by an amount $\Delta t$ corresponds to a phase displacement at frequency $2n$ of

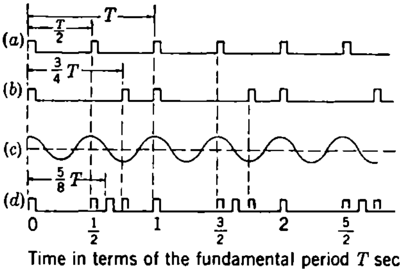


FIG. 11·1.—Space-modulation beacon communication. (*a*) A series of unmodulated pulses with a pulse-repetition frequency of $\frac{2}{T}$. (*b*) A series of modulated pulses similar to (*a*) but with alternate pulses delayed by an amount $\frac{T}{4}$ sec. (*c*) Second harmonic component of series of pulses in (*b*) (tone frequency). (*d*) A series of pulses similar to (*b*) but with alternate pulses space-modulated about the positions $\frac{5T}{8}$, $\frac{13T}{8}$, $\frac{21T}{8}$, etc.

$$2\pi \frac{\Delta t}{(T/2)} = \frac{4\pi \Delta t}{T}.$$

When the sets of pulses are symmetrically interlaced to give a regularly spaced set of pulses as in (*a*) of Fig. 11·1, $\Delta t$ equals $T/2$ and the phase displacement is $2\pi$. The two partial oscillations are then in phase and give a maximum resultant. When the spacing is $3T/4$, as in (*b*) of Fig. 11·1(*b*), the phase difference is $3\pi$ and the two oscillations are just out of phase and cancel to give a zero resultant. Obviously the general expression for the square of the resultant is given by $R^2 = 2 - 2 \cos 4\pi\Delta t/T$, which oscillates between 4 and 0. Its value as a function of $\Delta t$ is shown in Fig. 11·2.

If alternate pulses are moved slowly from position $T/2$ to position $3T/4$ [as in (*b*) and (*d*) of Fig. 11·1], the intensity of the tone of frequency $2n$ is gradually decreased to zero, according to the curve in Fig. 11·2. By abruptly shifting the alternate pulse position between $T/2$ and $3T/4$, therefore, the tone $2/T$ can be keyed on or off without changing the average working rate of the transmitter.

Now if, instead of varying between the extreme positions $T/2$ and $3T/4$, the pulse is moved between two positions closer to the center pulse position at $5T/8$, as in (*b*) and (*c*) of Fig. 11·3, alternate intervals of higher and lower intensity tone will result; and the higher intensity portions may be heard as dots or dashes as the case may be (see Fig. 11·3). If the amplitude of the pulse-position modulation decreases to zero (no movement of the pulse from its center position) a steady tone of half the maximum intensity is heard as in (*d*) of Fig. 11·3.

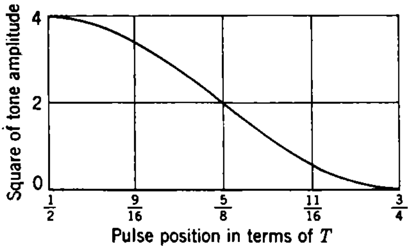


FIG. 11·2.—Tone-amplitude–pulse-position relationship. As the position of alternate pulses in Fig. 11·1 (*d*) varies from position $\frac{T}{2}$ to position $\frac{3T}{4}$, the square of the amplitude of the second harmonic varies according to the curve shown.

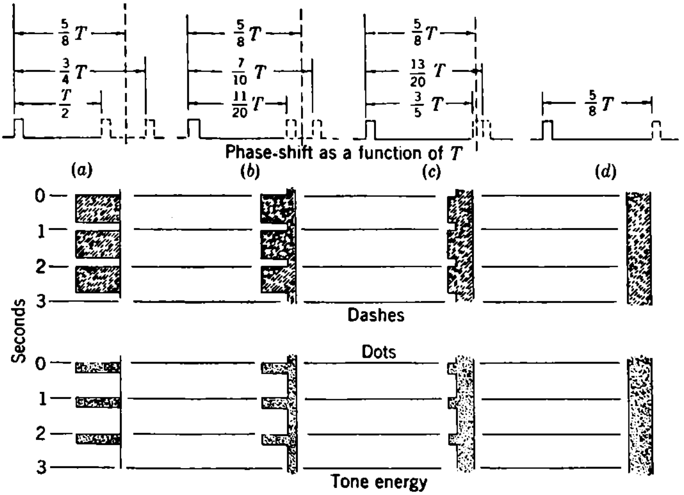


FIG. 11·3.—Phase-shift–tone relationship with square-wave modulation. Dots and dashes relative to background tone are shown for four different modulations. The pulses are shown modulated with square waves of ¾ sec duration spaced ¼ sec in the center part of the diagram and with the reverse timing in the lower section.

Course information for aircraft may be transmitted by this method. A series of dashes may indicate that the airplane is to the left of the course, a series of dots may mean to the right of the course, and a steady tone may mean on course. Like other methods for interpretation by an operator, this method can be used for continuous relative control; that is, instead of transmitting only the commands "left" or "right," an indication of "how far left" or "how far right" can be given. The amplitude of the dots or dashes relative to the background tone is then an indication of how far to the left or right of course the airplane is flying. The aural indications of this system are similar to aural indications received from the widely used A—N radio-beacon navigation system, and may, in fact, be identical with them. Space modulation of this type was used for communication in the Oboe beacon bombing system (see Chap. 1). With this method of modulation there is no variation at the modulation frequency of the average transmitted power.

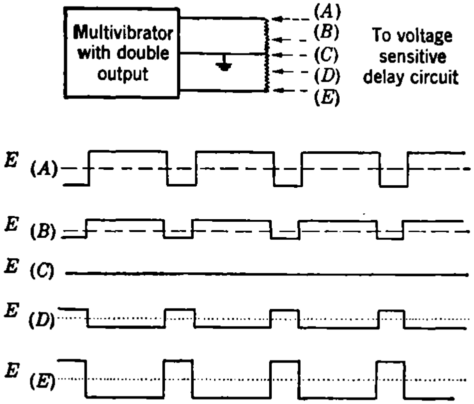


FIG. 11·4.—Continuous modulation of pulse position. A double-output multivibrator is connected to a center-tapped variable resistor with an adjustable arm. As the arm is moved from position (*A*) through (*B*), (*C*), and (*D*) to (*E*) the voltages at the arm are indicated by the waveforms $E_A$, $E_B$, $E_C$, $E_D$, and $E_E$, respectively. These voltages when fed to a voltage-sensitive delay circuit may be used to obtain space modulation of pulses.

The pulse position can be modulated in the way described above by manual control of the timing of alternate pulses through a voltage-sensitive delay circuit, as shown in Fig. 11·4. More often, the modulation is applied automatically by a circuit. For instance, the waveforms (*a*) and (*e*) of Fig. 11·4 can be applied to a double-triode mixer, the output voltage of which depends on the biases of the triode grids. The bias voltages may be obtained from the output of a double-gate range-error-indicating circuit.[1]

---

[1] For a discussion of such circuits, see *Waveforms*, Vol. 19, Chap. 14, and *Electronic Time Measurements*, Vol. 20, Radiation Laboratory Series.

**11·4. Pulse-width Modulation.**—Code letters can be sent and received by having the modulating signals vary the duration of the transmitted pulses if a pulse-width discriminator is followed by an audio amplifier at the receiver output. The receiver must, of course, be one that preserves the relative width of narrow and wide pulses.

By modulating the pulse width around an intermediate value in the manner shown for space modulation in Fig. 11·1, continuous relative control commands can be transmitted. For this purpose, the output voltage of the pulse-width discriminator in the receiver should be proportional to the pulse width over the region of modulation. This sort of modulation does produce a variation of the average transmitter power at the modulation frequency.

## METHODS OF INTERPRETATION BY AN OPERATOR

**11·5. Visual Methods.**—If an interrogator is searchlighting a beacon, the beacon reply is steadily visible on the interrogator scope. Now, if the beacon is keyed on and off according to some code, the code can be interpreted readily by an operator at the interrogator. The code may be simple, as an A—N code to give left-right directions, or it may be the Morse code, enabling the transmission of complex data. The keying of the beacon may be manual or automatic. For this method to operate satisfactorily, the beacon response must be steady.

It is possible and, indeed, preferable to produce a code by keying the width of the response pulse from one value to another. Because the pulses are usually short, a delayed fast sweep is often required at the receiving station for visual interpretation. The system must preserve the relative width of pulses.

**11·6. Aural Methods.**—In a system in which the beacon replies off frequency, the beacon response normally appears on the interrogator-receiver display without radar ground echoes. The initial interrogator pulse can also be gated out. Then, if an audio amplifier and speaker are connected to the output of the interrogator-receiver the pulse-repetition frequency $f$ will be heard only when a beacon responds. Keying the beacon on and off would then be a method of transmitting code. If more than one beacon responds, the audio-amplifier input can be range-gated so that beacons at different ranges can be distinguished.

If it is desired not to take the beacon completely off the air while keying, the response-repetition rate can be keyed by interrupting the beacon response at a low rate $f_1 \leq f/2$ when the key is pressed. The audio amplifier at the interrogator may then have a low-pass filter to pass $f_1$ but not $f$; the audio output then appears to be keyed on and off. Rejection of $f$ removes the initial pulse and ground echoes. Alternatively, the width of the reply pulses can be changed and a pulse-width discriminator inserted in the audio channel.

## METHODS FOR AUTOMATIC INTERPRETATION

**11·7. Pulse-repetition-frequency Systems.**—Signals in the form of pulses with a modulated repetition frequency can be interpreted automatically by filtering out all except the fundamental frequency component of the received pulses. The fundamental frequency component is then applied to a frequency discriminator.[1] The output voltage can be amplified and applied directly to the controlled element, or it can be applied through a servomechanism.

**11·8. Pulse-width Systems.**—Circuits for automatically interpreting coding signals based on pulse width have been described in Secs. 9·1 to 9·7. The same basic principles can be used for automatic interpretation of control signals using pulse width. One method is to convert the signals into sawtooth voltages (the peak amplitude of which is therefore proportional to the duration); to rectify and filter the sawtooth voltage, obtaining a d-c voltage proportional to signal width; and to apply this voltage to the controlled element either directly or through a servomechanism. Interference from unwanted signals may be prevented by using some additional form of coding—for instance, by having the control signals transmitted at a repetition frequency which differs from that of interfering signals and using a pulse-repetition-frequency discriminator.

**11·9. Multiple-pulse Systems.**—We have already considered (in Sec. 9·9) circuits for interpreting multiple-pulse identification signals automatically. Multiple-pulse signals can also be used in transmission of control signals for automatic interpretation. If a signal contains $N$ pulses, there are $N - 1$ spaces and each can be used to control a parameter. The method of measuring the pulse spacings is similar to that described in Sec. 9·10 for measuring the interval between recurring pulses. The first pulse starts a first linear sweep; the second pulse causes a first capacitor to be charged to the voltage of the first linear sweep at that instant, and so on. The capacitor voltages are then used directly or through mechanisms. It is necessary to sort out the pulses in the order of their arrival and to apply each at a different point; this is done by means of gated amplifiers switched on by pulses associated with each linear sweep. The first pulse may be identified from interfering signals by using some form of code—for instance, a pulse-width or a double-pulse code.[2]

---

[1] Circuits that respond to particular pulse-repetition rates have been discussed in Sec. 9·10 of this volume.

[2] For a description of complete multiple-pulse systems see *Waveforms*, Vol. 19, Chap. 10, and *Electronic Time Measurements*, Vol. 20, Chap. 10, Radiation Laboratory Series.

Use of each code space allows control of a parameter over a limited range, but a parameter $\theta$ can be controlled over an unlimited range if both sin $\theta$ and cos $\theta$ are transmitted, each by a different code space. However, unless special precautions are taken, there will be an ambiguity in $\theta$ of $2n\pi$, where $n$ is any integer.

In multiple-pulse systems, all pulses need not be transmitted on the same radio frequency. Greater privacy and automatic separation result from using separate frequencies.

## SPEECH TRANSMISSION ON BEACON SYSTEMS

**11·10. Pulsed Voice Systems.**[1]—As was pointed out previously, it is necessary to add special equipment to both interrogator and beacon to allow voice communication in the interrogator-beacon system.

It was shown in Sec. 5·18 that channels wide enough for voice communication may be obtained even with low pulse-repetition frequencies, if a sufficiently wide r-f bandwidth is used. In such systems more than one element of information must be transmitted by each pulse. Systems of this nature are clearly more difficult to design than systems in which only one element of information is transmitted by each pulse; as far as we are aware no such system has yet been proposed. Our discussion will be limited, therefore, to systems in which the pulse-repetition frequency is used as the carrier frequency for the voice modulation, and in which the pulse-repetition frequency, consequently, is 4000 cps or more. It is possible, however, to send some pulses on one frequency and the remainder on another, and to distinguish between the replies to the pulses. If this is done, a lower pulse-repetition frequency may be used for the beacon range and azimuth data, and only replies to the lower pulse-repetition frequency displayed on the cathode-ray tube. The severe restrictions on range resulting from high pulse-repetition frequencies can thus be removed.

It is well known that a set of uniformly spaced pulses having a pulse-repetition frequency $F_0$ can be represented or replaced by a Fourier series of terms involving frequencies of 0 (the d-c term), $F_0$, $2F_0$, $3F_0$, etc. When the amplitude of the pulses is modulated at a frequency $F_m$, each of the original frequencies $NF_0$ in the expression for the unmodulated pulses is replaced by a triplet of frequencies $NF_0 - F_m$, $NF_0$, $NF_0 + F_m$, as with ordinary amplitude modulation of continuous waves. This is not surprising since the component terms of the Fourier expansion are continuous. More careful examination shows that when the modulating envelope is considered actually to modify the shape of the top of the pulses the two sideband frequencies $NF_0 - F_m$ and $NF_0 + F_m$ will be of equal amplitude. When, however, the pulses are assumed to be rec-

---

[1] Secs. 11·10 and 11·11 by M. D. O'Day.

tangular with heights corresponding to the position of their centers on the curve of modulation, the sidebands will, in general, not be of precisely equal amplitude. The relative amplitudes will then be given by the familiar expression $\sin \pi\tau F/\pi\tau F$ where $F$ is the actual frequency and $\tau$ is the duration of the pulse. These points are discussed more fully in Appendix A.

A characteristic of pulse modulation that tends to limit its usefulness is an ambiguity with respect to the frequency of the modulation. If regularly spaced rectangular pulses having a pulse-repetition frequency $F_0$ are modulated in amplitude with a frequency $F_m$, the resulting set of pulses is *entirely indistinguishable* from other sets of modulated pulses where the modulation frequencies are $NF_0 + F_m$, $N$ taking different integral values. These equivalent modulations must, of course, have the appropriate values of amplitude and phase. To see that this is true, consider a set of rectangular pulses centered at times 0, $1/F_0$, $2/F_0$, $3/F_0$ . . . with their heights given by the expression

$$A = A_0[1 + p \cos (2\pi F_m t + \phi)],$$

in which $A_0$ is the average height of the pulses, $F_m$ is the frequency of modulation, $p$ is a fraction expressing the percentage of modulation, and $\phi$ is a phase factor put in for the sake of generality.

The values of the cosine term for the successive pulses are

$\cos \phi$,
$\cos (2\pi F_m/F_0 + \phi)$,
$\cos (4\pi F_m/F_0 + \phi)$, etc.

If now, for example, we substitute $NF_0 - F_m$ in place of $F_m$ and take the phase to be $-\phi$, the successive values will be

$$\cos (-\phi) = \cos \phi$$

$$\begin{aligned}\cos\left[\frac{2\pi(NF_0 - F_m)}{F_0} - \phi\right] &= \cos\left[2\pi\left(N - \frac{F_m}{F_0}\right) - \phi\right]\\ &= \cos\left[-\left(\frac{2\pi F_m}{F_0} + \phi\right)\right] = \cos\left(\frac{2\pi F_m}{F_0} + \phi\right),\\ \cos\left[\frac{4\pi\,(NF_0 - F_m)}{F_0} - \phi\right] &= \cos\left[4\pi\left(N - \frac{F_m}{F_0}\right) - \phi\right]\\ &= \cos\left[-\left(\frac{4\pi F_m}{F_0} + \phi\right)\right] = \cos\left(\frac{4\pi F_m}{F_0} + \phi\right), \qquad \text{etc.}\end{aligned}$$

Comparison of this set of values of the cosine term with the former set shows that they are identical. Thus, there is a modulation of frequency $NF_0 - F_m$ that gives a set of pulses identical with those given by modulation at frequency $F_m$. The same argument is obviously applicable for all the frequencies of modulation given by $NF_0 + F_m$.

The practical consequence is that the modulation frequencies used with pulses should be limited to values less than half the pulse-repetition frequency, since any modulation frequency $F_m$ that lies between $F_0/2$ and $F_0$ can give an apparent modulation at $F_0 - F_m$ which will be less than $F_0/2$ by the same amount that $F_m$ is greater than $F_0/2$. Similarly the receivers for modulated pulses should contain filters to eliminate modulation frequencies greater than $F_0/2$. Thus, if voice frequencies up to 2000 cps are to be used for modulation, the pulse-repetition frequency must exceed 4000 cps.

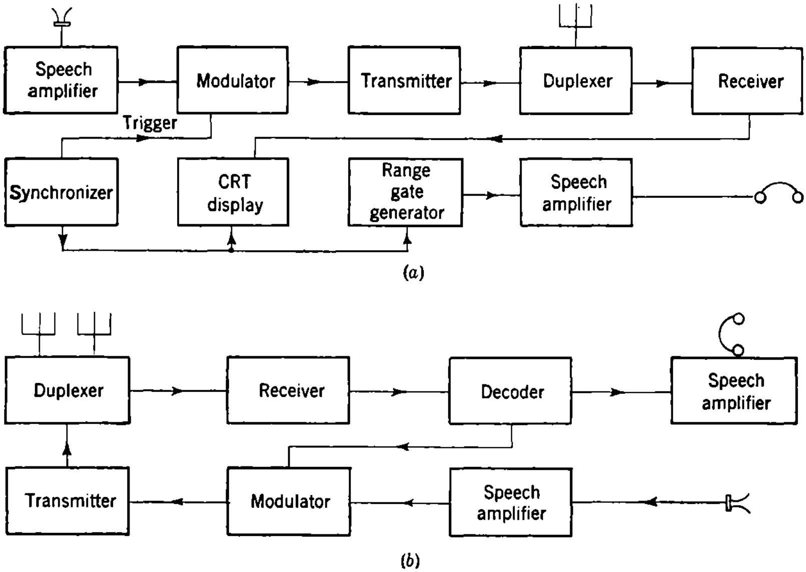


FIG. 11·5.—A two-way beacon speech communication system. (a) Interrogator. (b) Beacon.

## 11·11. Types of Modulation.

**11·11. Types of Modulation.**—Perhaps the simplest type of modulation which can be employed is frequency modulation of the pulse-repetition frequency. When this method is used, the beacon is equipped with an f-m discriminator which converts frequency modulation into amplitude modulation. The system is then restricted to the use of one interrogator at a time for each beacon.

Other types of modulation which can be employed are double-pulse modulation, in which the spacing between a pair of pulses is used as the parameter to be modulated, pulse-width modulation, and pulse-amplitude modulation. In general, such modulation is detected by transforming it to amplitude-modulated pulses so that most of the important relationships can be brought out by consideration of amplitude modulation.

Figure 11·5 shows a block diagram of a two-way beacon speech-communication system.

**11·12. "Talking Rebecca-Eureka."**[1]—The simplest voice-communication beacon system that has been worked out, and the only one that seems to have been used in the field, is that employed in the "talking Rebecca-Eureka" system. The Rebecca-Eureka system has been discussed in Secs. 1·7 and 4·10; it consists of an airborne interrogator with lobe switching for azimuth determination and a very light portable ground beacon with an omnidirectional antenna. No interrogation coding is used, and response coding is by manual pulse-width keying.

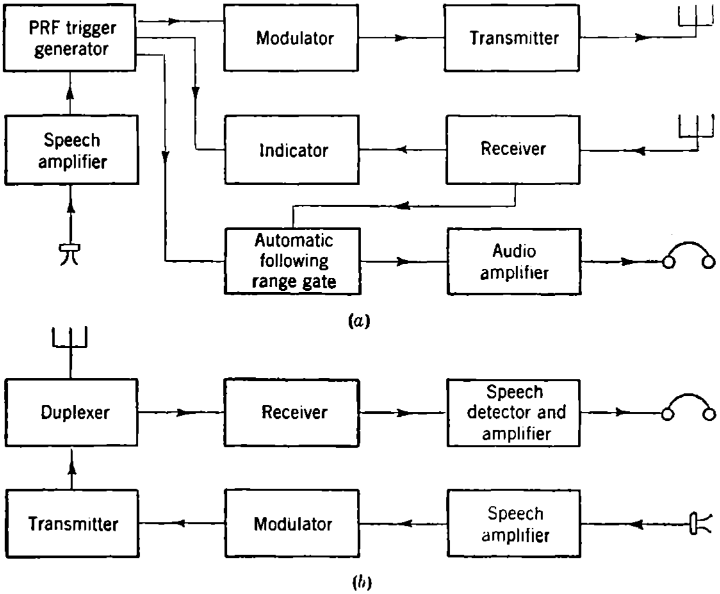


FIG. 11·6.—Block diagram of talking Rebecca-Eureka. (a) Interrogator. (b) Beacon.

The useful range of the system is short and the interrogation is essentially continuous. As a result, it has proved possible to introduce two-way voice communication into the system without impairing the performance of the beacon except by the inevitable reduction in traffic capacity consequent on its use as part of a communication system. Figure 11·6 shows a block diagram of the talking Rebecca-Eureka.

When a Rebecca-Eureka system is to be used for voice communication, the pulse-repetition frequency of the interrogator is increased to about 5000 cps; the range of the system is thus limited to about 15 miles.

---

[1] By A. Roberts.

Responses from more distant beacons appear on incorrect sweeps. Ambiguities in range may be resolved by momentarily changing the pulse-repetition frequency. Modulation is introduced in the interrogation link by frequency-modulating the pulse-repetition frequency. The beacon is equipped with a discriminator tuned to the mean pulse-repetition frequency; its output is an audio signal corresponding to the voice modulation.

In the response link, modulation is accomplished by varying the pulse width of the response. No range error or variation is thus introduced into the normal beacon response, which appears as before on the interrogator scope. The interrogator receiver is equipped with an integrating circuit which converts the pulse-width modulation into an amplitude-modulated audio output. Range-gating of the beacon response is used both to reduce ignition and other interference and to select the correct beacon response.

Fig. 11·7.—Double-pulse modulation. The position of the modulation pulse *M* varies periodically with respect to the reference pulse *R*.

**11·13. Double-pulse Modulation.**[1]—An easy method of modulation is to send out paired pulses, as shown in Fig. 11·7. One pulse acts as a reference pulse, and the second pulse is called the modulation pulse. It is not necessary that these pulses be sent on the same carrier frequency; the reference pulse may be that of a radar, and the modulation pulse may be produced by an interrogator associated with it.[2]

## Voice-modulation Systems to Work with Radars

By M. D. O'Day and R. A. Minzner

The following hypothetical paired-pulse communication system has been suggested for use in conjunction with ground radar. The reference pulse is that of the ground radar. Although the modulation pulse can be on any carrier frequency, there are certain advantages in having this frequency low enough to make omnidirectional antennas effective. The system is most useful when operated with a ground-interrogator–airborne-beacon system; it might also be used with airborne interrogators working with portable ground beacons.

---

[1] By M. D. O'Day.

[2] It may be possible to combine various types of modulation to form a more secure communication system and also to improve the signal-to-noise ratio. One method might be to combine, in the audio part of the receiver, modulation envelopes sent both by pulse-repetition frequency modulation and double-pulse modulation. Circuits would be introduced which would not respond to any signals except those which are present at the same time and in the same phase with both types of modulation.

The system will be illustrated by two examples:

1. A short-range interrogator with a high pulse-repetition frequency.
2. Other radars with pulse-repetition frequencies too low for voice communication.

**11·14. Communication with a Radar of Very High Pulse-repetition Frequency.**—Let us assume that we have a radar with a pulse-repetition frequency in excess of 4000 cps. Associated with it is an interrogator which transmits a modulation pulse shortly after each radar pulse. The delay time of the modulation pulse can be varied with respect to the reference pulses. At the receiving end, the time between the reference pulse and the modulation pulse is integrated and converted into an audio signal.

A considerable degree of security can be provided by the use of separate frequencies for the two pulses. Suppose the reference pulse has a random "jitter" which is several times greater than the total excursion of the modulation pulse with respect to the reference pulse. At the beacon end, then, it would be necessary to receive the reference pulses in order to decode the information, because the random jitter of the reference pulse with respect to any fixed repetition period would correspond to a high noise level.

A second feature of this system is its ability to reject random pulses that may cause interference. Suppose that the excursion of the modulation pulse around its average value is only 20 per cent of the spacing between the two pulses. Decoder circuits which will not recognize any pulses that do not have a spacing corresponding to this range can then be used.

The beacon transmitter and modulator are somewhat simpler. It is unnecessary to transmit the reference pulses back to the interrogator since they are already available there. The beacon replies are varied periodically in range around an average value which is somewhat greater than the range of the equipment as measured by the radar. The range measurement is corrected for the delay in the beacon reply. The demodulation at the interrogator is similar to that at the beacon. Range-gating is used to select the correct beacon response for the input to the audio channel.

The operational procedure at the interrogator is simple. The operator searchlights the beacon and obtains the beacon reply on the indicator. He then adjusts a range gate which brackets the beacon reply (see Fig. 11·8). The system is then ready for operation. The interrogator operator can talk to the beacon operator and he will receive only replies which are at the same range as the beacon. Azimuth resolution depends upon the interrogator beamwidth. As a consequence, there is usually no ambiguity as to which target is talking to him.

This feature of the system has two advantages; it identifies the conversation as orginating from a particular target, and it eliminates a great deal of interference and noise. Only interference which enters the range gate is received. The identification of a particular signal as corresponding to that of a selected target on the radar indicator can be of great advantage in ground control of aircraft flying under conditions of poor visibility.

Figure 11·9 shows a block diagram of an experimental beacon used in a communication system of this type. The beacon is interrogated simultaneously by a 10-cm radar and a 500-Mc/sec interrogator-responsor. It replies at about 500 Mc/sec. Double-pulse modulation is used in both interrogation and reply links.

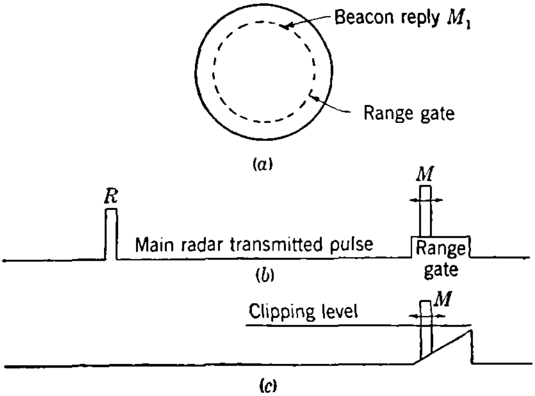
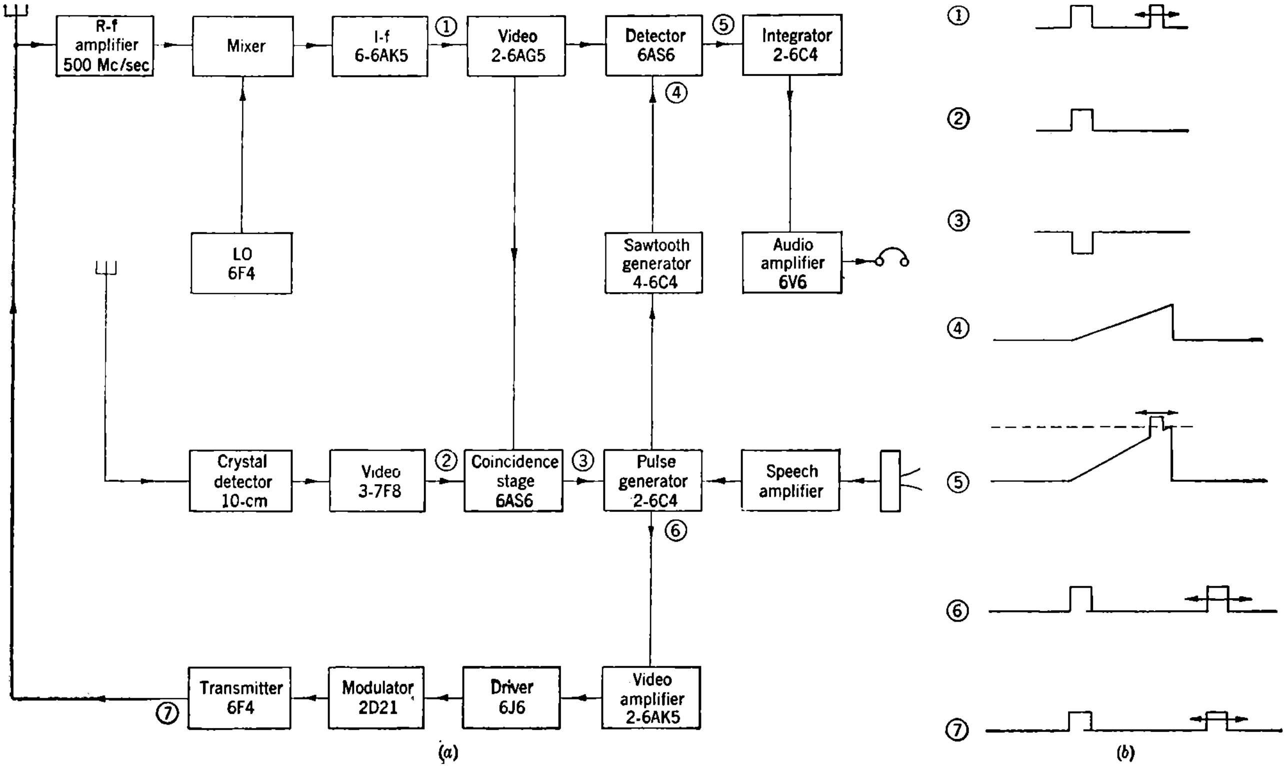


FIG. 11·8.—Operation of range-gated beacon communication system. (a), (b) The operator brackets the beacon reply with a range gate. (c) The range gate triggers the sawtooth gate on which rides the pulse $M$. The fraction of $M$ above the clipping level changes as it changes its position with respect to the range gate. NOTE: This will work with self-synchronous systems since the range gate is initiated by the main radar pulse.

**11·15. Voice Communication with Low-repetition-rate Radars.**—Because powerful radars do not have high repetition rates, the system just described is greatly restricted in its use. It has little advantage over the simpler talking Rebecca-Eureka. It is not difficult to modify it for use with low-repetition-rate radars, however, retaining most of its advantages. Modification can be accomplished readily because the intelligence does not have to be contained in the reference pulse but may be largely in the modulation pulse. It is necessary then to transmit several modulation pulses corresponding to each reference pulse. If there is a random jitter in the radar, these trains of pulses will jitter with the reference pulses. The system is illustrated in Fig. 11·10. For simplicity, we have taken a case in which four modulation pulses are required for each radar pulse. In the beacon, each reference pulse sets up several



(a) (b)

FIG. 11·9.—(a) Block diagram of an experimental coincident crossband voice communication beacon. Double-pulse modulation is used in both interrogation and reply links. (b) Waveforms at numbered points. (1) Received double-pulse interrogation on 500 Mc/sec. (2) Received interrogation on 10-cm—a single pulse coincident with the first pulse of the 500-Mc/sec pair. (3) Coincidence stage output trigger for beacon range reply pulse. (4) Sawtooth voltage sweep used for demodulation of 500-Mc/sec. interrogation. (5) Modulation detector output waveform. The amount of the modulation pulse above the clipping level varies with the pulse spacing. (6) Reply pulse pair, generated by the pulse generator. The pulse spacing varies with the applied audio voltage. (7) Beacon reply at 500 Mc/sec.

gates (four in Fig. 11·10 instead of one). The circuits are necessarily more complicated than in the first system described; some adjustment must be available to accommodate radars of differing repetition rates. In other respects, however, the system acts just as in the case in which the radar has a high recurrence rate.

It is to be noted, however, that in order to keep replies to the modulation pulses from showing on the indicator, some type of response coding should be used to differentiate them from the beacon reply to the radar pulses. Such response coding may be pulse-width coding, or by reply on a different frequency. If a different frequency is used, the beacon must have two transmitters, and the interrogator-beacon system has been complicated to the extent of adding an additional transmitter and receiver at each end of the system. If pulse-width coding is used, replies to the radar pulses alone can be allowed to appear on the display, and the demodulation of all the beacon reply signals can be carried out as in the previous example. Figure 11·11 shows a pulse diagram of such a system.

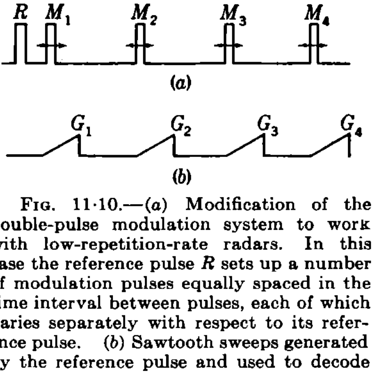


FIG. 11·10.—(a) Modification of the double-pulse modulation system to work with low-repetition-rate radars. In this case the reference pulse $R$ sets up a number of modulation pulses equally spaced in the time interval between pulses, each of which varies separately with respect to its reference pulse. (b) Sawtooth sweeps generated by the reference pulse and used to decode the modulation.

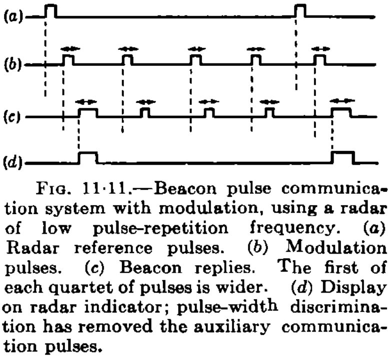


FIG. 11·11.—Beacon pulse communication system with modulation, using a radar of low pulse-repetition frequency. (a) Radar reference pulses. (b) Modulation pulses. (c) Beacon replies. The first of each quartet of pulses is wider. (d) Display on radar indicator; pulse-width discrimination has removed the auxiliary communication pulses.

**11·16. Use of More Complex Modulation Envelopes.**—In the preceding sections systems using only the characteristics of simple rectangular pulses—amplitude, duration, and spacing—have been described. It is theoretically possible to pack an almost unlimited amount of information into a pulse—for instance, by having the r-f amplitude or frequency envelope take on very complicated forms. One practical limitation is in the bandwidth required; another is that high-power magnetrons do not readily permit amplitude or frequency modulation. The transmission of data more complex than any described above, however, is possible. For instance, one line of a television transmission may be regarded as a pulse which contains data expressing the value of a function at every point on a line of finite length.

# CHAPTER 12

# MODULATORS

BY P. A. DE PAOLO, J. J. G. MCCUE, K. R. MORE, AND J. C. REED

**12·1. General Considerations.**[1]—The beacon modulator is essentially a device for supplying a voltage pulse to a transmitting tube. Its function, therefore, is the same as that of a radar modulator and it has much the same general design. It differs from the radar modulator, however, in three important respects:

1. When range coding is used, the interval between pulses may be only about 10 μsec, or even less in some cases.
2. Usually there is no fixed repetition rate; the duty ratio fluctuates with time.
3. All pulses emitted by the transmitting tube must have the same frequency; the voltage pulse delivered by the modulator must remain constant in spite of variations in duty ratio and line voltage.

The small interval between pulses may make it difficult to use gas-filled tubes as switches. The time required for recombination of the ions in a gas-filled tube is of the order of 50 μsec or more. Furthermore, a gas-filled tube must be used in conjunction with a pulse-forming network which cannot be recharged very quickly. It will be shown below that these difficulties do not entirely exclude gas-filled tubes from use in beacon modulators, but they do account for the fact that the modulator for a range-coded beacon usually consists of a hard-tube switch driven by a blocking oscillator.

The greatest differences between beacon and radar modulators arise from the need for frequency stability in a beacon. The two principal causes of frequency instability are spontaneous frequency shifting in magnetrons and dependence of transmitter frequency on modulator output.

Spontaneous frequency shifting (often incorrectly called "moding" or "double moding") is a discontinuous change in the output power and frequency of a magnetron when there is no intentional change in the pulse applied to the magnetron. The magnetron oscillates at one fre

---

[1] By J. J. G. McCue and P. A. de Paolo.

quency during some of the pulses and at another during the rest of the pulses. Frequency shifting is much more likely to cause trouble in beacons than in radars for two reasons. The use of a reactive r-f load to pull the magnetron to a spot frequency increases the tendency of a magnetron to shift frequency spontaneously, and the presence of frequency shifting in a beacon may make the identification of a range code difficult or impossible. In a radar, on the other hand, it simply lowers the effective pulse-repetition frequency. Figure 12·1 shows two beacon signals appearing on the screen of a radar.

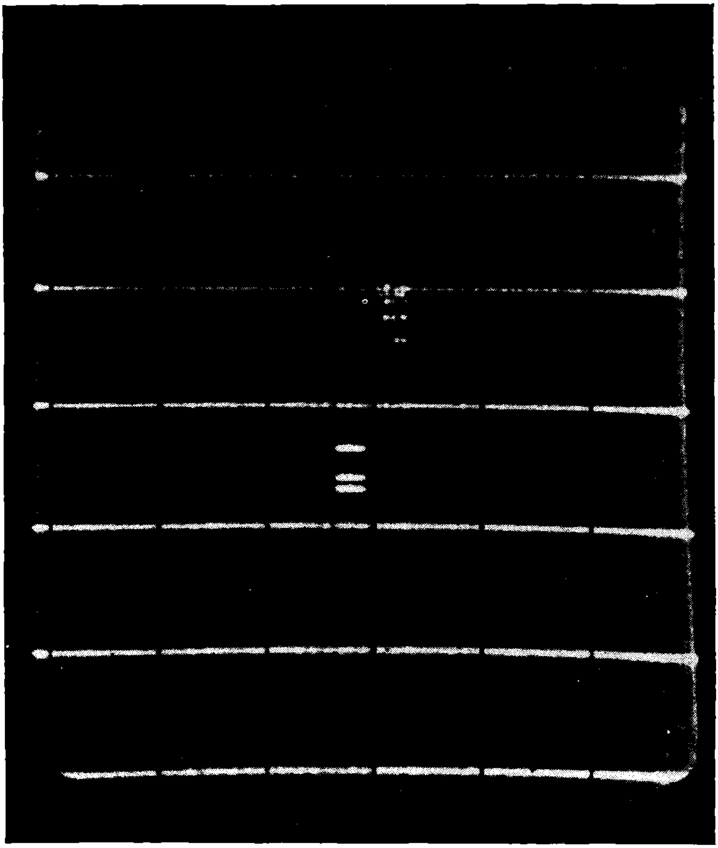


FIG. 12·1.—Effect of spontaneous frequency shifting on the appearance of a beacon response. In the B-scope photo, the beacon whose code is 2-1 shows a normal response. The response of the other beacon, whose code is 1-2-3, is erratic, with many pulses missing, so that the code is difficult to read and the response weak.

The beacon on the right exhibits spontaneous frequency shifting; that on the left does not. Spontaneous frequency shifting often can be prevented by making the time of rise of the voltage pulse applied to the magnetron sufficiently large—for example, 0.2 μsec (see Secs. 13·3 and 13·4)—and can be inhibited by running the magnetron at a carefully chosen operating point.

If possible, the transmitting tube should be so operated as to have a low "pushing figure"—that is, changes in the current pulse in the transmitter should cause only a slight change in the frequency of the r-f output. Once the tube has been chosen, preventing it from being "pushed" away from its assigned frequency is essentially a matter of regulation. The modulator and power supply should be designed so that the pulse applied to the transmitting tube does not vary with line voltage or duty ratio or from pulse to pulse in a code group.

The pulse applied by the modulator to the transmitting tube may be formed in the coder and amplified and shaped by the modulator itself. The advantages of this method are that the coder design need concern itself only with the spacing of the code pips, without regard to their shape, and that it avoids the difficulty of maintaining the shape of short pulses passing through several stages of amplification. In either case, the pulse may be formed in one of the ways described below.

## PULSE-FORMING CIRCUITS

BY J. J. G. MCCUE AND P. A. DE PAOLO

The principal circuits used for pulse formation in beacons are multivibrators, line-controlled blocking oscillators, and pulse-forming networks used in conjunction with gas-filled tubes.

### HARD-TUBE PULSE-FORMING CIRCUITS

**12·2. Multivibrators.**—The use of multivibrators for pulse formation is discussed in *Waveforms*, Vol. 19, Chap. 5, Radiation Laboratory Series. It is enough to note here that if a multivibrator is to generate short rectangular pulses, the time constants of the circuit and also the interelectrode capacitances of the vacuum tubes must be kept small, and that tubes with high mutual conductance give steeply rising pulses. Figure 12·2 shows a circuit that has been used to produce short pulses of adjustable length.

The chief advantages of multivibrators for rectangular pulse formation are

1. Very flat-topped pulses are obtainable.
2. No transformer is needed.
3. Long pulses are easily obtained.
4. Low peak currents permit continuous operation with very short intervals between pulses.
5. Adjustable pulse length is easily obtained.

Their disadvantages are low pulse-power output combined with high average power input.

The multivibrator can be made to give rectangular pulses as short as 1 μsec. For pulses less than 10 μsec long, however, a blocking oscillator is usually preferable to a multivibrator because of its superior efficiency.

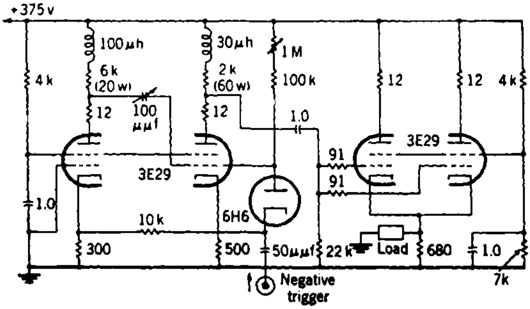


FIG. 12·2.—Multivibrator and cathode follower for 1- to 100-μsec pulses. The pulse length is controlled by varying the 100-μf condenser and the 1-M resistor.

For pulses less than 2 μsec long, it is very difficult to make a multivibrator work at all without using tubes that consume a great deal of power.

**12·3. Blocking Oscillators.**—The form of blocking oscillator, or regenerative pulser, which has been found most useful in beacon modulators is shown in Fig. 12·3. This circuit, when set into operation by a trigger pulse of noncritical shape, produces a rectangular pulse of predetermined width, at a power level that may be very high. For example, 25-kw pulses, 1 μsec long, are obtainable from a 3E29 tube in a circuit that functions on a 100-volt trigger obtainable from a 6SN7 cathode follower. Such a circuit can, of course, be used to drive a transmitting tube; but operation that is less critical, especially with regard to the load impedance, can be obtained by using a low-power blocking oscillator to drive a switch tube, which in turn drives the transmitting tube.

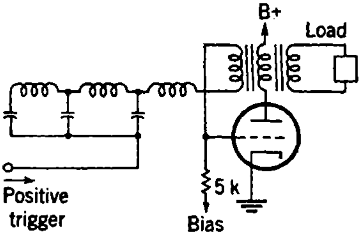


FIG. 12·3.—Series-triggered line-controlled blocking oscillator.

The theory of the blocking oscillator is discussed in *Pulse Generators*, Vol. 5, Sec. 14·2, and in *Waveforms*, Vol. 19, Chap. 6 of this series. In brief, the blocking-oscillator grid potential is raised above cutoff by a trigger pulse and regeneration drives the grid further positive. The plate current rises rapidly at first and the plate voltage drops. Then, if the circuit satisfies certain conditions we will not discuss here, the operating point of the tube changes in such a way as to maintain a constant voltage in the output winding of the transformer. Meanwhile, the current in the grid circuit charges the delay line. After a time depending on the characteristics of the delay line, the voltage across the terminals of the delay line changes abruptly and causes a sharp drop in the grid potential. The corresponding drop in plate current drives the grid beyond cutoff by regenerative action. The charge on the delay line leaks off through the grid resistor, and the circuit is ready to be triggered again.

The time required for the grid potential to return, after the pulse, to a value such that the trigger pulse can set off a new pulse, is called the "recovery time." It is a particularly important parameter in beacon modulators because it sets a lower limit on the spacing between pips in a range code. The recovery time of a blocking oscillator depends on the nature of the trigger pulse and on the grid bias; it also depends on the grid resistor and the capacitance of the delay line. Since, for a given pulse duration, the capacitance of the delay line is inversely proportional to the square of the characteristic impedance of the line, a high-impedance line makes for a short recovery time.

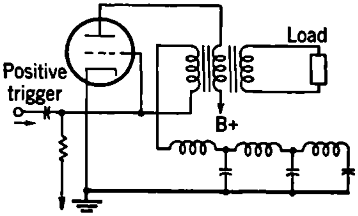


FIG. 12·4.—Parallel-triggered line-controlled blocking oscillator.

If the output pulse of the blocking oscillator is to be rectangular, the characteristic impedance of the delay line must be matched to the generator that charges it, which is the transformer feedback secondary winding in series with the impedance of the trigger source and the grid-cathode impedance of the tube. For this reason, it is well to use in the blocking oscillator a tube having a low amplification factor. Such a tube will require a small ratio (for example, 1-to-1) of primary turns to feedback secondary turns in the transformer. This implies a high impedance for the feedback winding and therefore a high impedance for the delay line.

It might be inferred that a trigger source with high internal impedance is desirable. This is not true, however, because a high-impedance trigger source would limit the grid current, and thereby limit the output power and impair the rectangular shape of the pulse. A cathode follower makes a good trigger source.

When the impedance of the trigger source is high, the circuit shown in Fig. 12·4 may be used. Here the grid current does not pass through the trigger source.

The chief advantages of the blocking oscillator are that it draws current only during the pulse and that it generates a high voltage in a low-impedance output. It is therefore a very efficient pulse generator compared with a multivibrator. For pulses less than 10 μsec long, it has few disadvantages in modulation applications; the top of the pulse can be made nearly flat, and the possible inconvenience arising from the need for a transformer and delay line is outweighed by the high power level of the output.

## LINE-TYPE PULSE-FORMING CIRCUITS

**12·4. Pulse-forming Networks and Switch Tubes.**—The line-type pulse-forming circuit, widely used in radar modulators, consists of a pulse-forming network like a lumped-constant transmission line, a low-impedance switch to discharge the network into the load, and means for recharging the network. In a beacon, the switch must be triggered by the coder output; the switch is therefore a gas-filled tube. An appropriate circuit is shown in Fig. 12·5. Here the charging circuit is simply the power supply and the resistor $R_c$ which, during the charging process, limits the current in the tube to a value that prevents maintenance of the discharge in the gas.

When the switch tube is fired, its resistance is so low that the charged network is connected across a load that consists mainly of the primary of the pulse transformer. If the characteristic impedance of the network matches the impedance of the load, the network discharges completely into the load, producing an approximately rectangular pulse the duration of which is determined by the design of the network. The magnitude of the voltage pulse across the primary of the transformer is half of the supply voltage minus the voltage drop in the gas-filled tube.

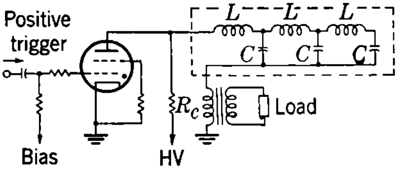


FIG. 12·5.—Pulse-forming network and gas-filled switch-tube modulator.

If the pulse-forming network in Fig. 12·5 is a lumped-constant transmission line as shown, then it will usually consist of from two to five sections, the inductance and capacitance of which are first computed from the desired characteristic impedance and pulse duration, and then altered empirically to give a suitable pulse shape across the transformer secondary. In general, the larger the number of sections, the flatter the pulse. Other forms of network are sometimes used.[1]

*Resistance Charging.*—When the network has discharged into the load, the resistor $R_c$ limits the current in the switch tube to such a low value

---

[1] The designer will find detailed information on pulse-forming networks in *Pulse Generators*, Vol. 5, Chap. 6, Radiation Laboratory Series. Gas-filled switch tubes are discussed in Chap. 8 of the same volume.

that the ions recombine, and the tube ceases to conduct after a time that depends on the circuit and on the nature of the gas in the tube; it is of the order of 10 to 100 μsec. The pulse-forming network then charges through the resistance $R_c$. Since $R_c$ is ordinarily large compared with the characteristic impedance of the network, the charging is very nearly exponential, with a time constant equal to $R_c$ times the total capacitance $C_N$ of the network. The blanking gate preceding the modulator should be of such length that the switch tube will not be triggered again until the network is sufficiently charged. After a time $4R_cC_N$ has elapsed, the network will be charged to 98 per cent of the supply voltage. Recharging can be hastened by placing an inductance in series with $R_c$; the effect is shown in Fig. 12·6.

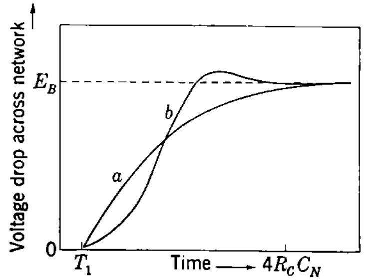


FIG. 12·6.—Resistance- and inductance-resistance charging. (a) Resistance-charging characteristic. (b) Inductance-resistance charging characteristic. $T_1$—recombination time of gas ions in switch tube.

A recently suggested method for eliminating the "dead" time due to exponential recharging is described in *Pulse Generators*, Vol. 5, Chap. 11, Radiation Laboratory Series. The network is first charged through the load when one switch tube is conducting, then discharged through the load while another switch tube conducts. By using two primaries in the transformer, output pulses that all have the same polarity can be obtained. The interval between pulses can be as small as the recombination time of the tubes.

When the spacing of the pulses has to be less than the recombination time of a gas-filled tube, a separate pulse-forming network and switch tube can be used for each pulse, all of the network-switch combinations being connected in parallel across the load. In this way, a closely spaced code group can be transmitted.[1] Such a circuit was briefly described in Sec. 10·10.

**12·5. Other Charging Methods.** *D-c Resonance Charging.*—Energy dissipation in the resistor $R_c$ reduces the efficiency of the line-type pulse-forming circuit. In interrogators, or in beacons designed to operate at fixed repetition rates, resistance charging of the pulse-forming network can be replaced by more efficient methods. These methods depend on substituting for the charging resistor $R_c$ an inductance $L_c$, which combines

---

[1] A circuit of this sort is discussed in detail in *Pulse Generators*, Vol. 5, Chap. 11, Radiation Laboratory Series, which also describes a circuit for getting still smaller spacing between pulses by means of an open-ended delay line used in conjunction with a single switch and pulse-forming network.

with the capacitance $C_N$ of the network to form a resonant circuit whose natural frequency is half the pulse repetition frequency. That is, $L_c$ is chosen so that $\pi\sqrt{L_cC_N} = T_R$, where $T_R$ is the interval between pulses. Because of damping, the first surge of current in the capacitor charges it to a little less than twice the supply voltage $E_B$, that is, to about $1.9E_B$. The switch tube is triggered at this time, and the network discharges a pulse of about $0.95E_B$ into the switch tube and transformer; then the cycle is repeated. The voltage across the pulse-forming network is shown as a function of time in Fig. 12·7 where the dotted line shows the oscillations that would occur if the switch tube were not fired. In the figure, the pulse duration is exaggerated to show the voltage step in the discharge. The recombination time, however, is not shown in the figure because, although it is usually longer than the pulse duration, it is usually too short to be noticeable on the scale of the figure when resonance charging is used.

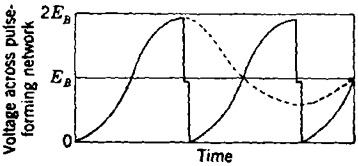


FIG. 12·7.—D-c resonance charging. The dotted line shows the damped oscillation which would occur if the switch tube were not fired at the peak of the charging cycle.

This charging method avoids most of the loss inherent in charging through a resistance and also permits the use of a lower power-supply voltage. It has the disadvantage of requiring a change in the inductance when the network capacitance $C_N$ is altered to change the pulse duration, or when the interval $T_R$ between pulses is changed. This disadvantage can be eliminated by placing a "hold-off" diode in series with the inductance $L_c$, so that the network cannot discharge through the inductance. Another reason for using a hold-off diode is that it permits the use of a smaller inductance. The hold-off-diode arrangement might be applied to a beacon operating at an irregular rate; this has not been attempted.

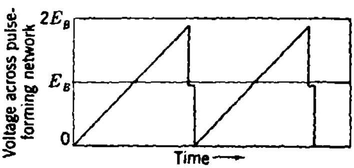


FIG. 12·8.—D-c straight-line charging.

*Straight-line Charging.*—An alternative method for inductance charging, which permits a change in the repetition rate, is that known as "straight-line charging." In this method a charging inductance about four or more times as large as that required for resonance charging is used. The mathematical analysis is somewhat involved,[1] but physically it is clear that the large inductance tends to keep constant the current in the charging circuit, and therefore the voltage across the network will change at a nearly constant rate. The network voltage as a function of time,

---

[1] See *Pulse Generators*, Vol. 5, Chap. 9, of this series.

after a steady state has been reached, is shown in Fig. 12·8. The fundamental difference between d-c resonance charging and straight-line charging is that in straight-line charging the current in the charging inductance is not zero when the network begins to recharge. Straight-line charging permits switching from one repetition rate to another. It has not, however, been used in beacons. Because of the steep slope of the charging curve at the instant of firing, straight-line charging is undesirable when the repetition frequency is subject to short-time fluctuations. In such cases, d-c resonance charging, perhaps with a diode, is indicated.

*A-c Resonance Charging.*—The pulse-forming network can be charged through an inductance connected directly to the power transformer, provided the power-supply frequency is constant and bears an integral relationship to the repetition frequency. This is called "a-c resonance charging"; it has been used successfully on certain large fixed radar installations, where the alternating current for charging is obtained from a generator run by a synchronous motor. It is discussed at some length in *Pulse Generators,* Vol. 5, Chap. 9, Radiation Laboratory Series. It has not been used in beacons.

**12·6. Summary.**—The advantage of a line-type pulse-forming circuit is its large output power with simple circuit design. Its disadvantages are that it produces rather poorly shaped pulses and that it will not function at high repetition rates nor even, in some cases, at variable rates. Furthermore, it introduces a delay that may be as large as 1 μsec, which is subject to change when the gas-filled tube is changed. The circuit has the further disadvantage that a 10 per cent change in supply voltage causes a change in output power of about 20 per cent, because the power output is proportional to the square of the supply voltage.

## PULSE FORMATION IN MODULATORS

The blocking oscillator and the line-type pulser, which are the pulse generators usually employed in beacon modulators, can either generate the pulse at the power level required by the transmitter or they can generate low-power pulses that drive the grid of a hard-tube switch. In the interests of simplicity and efficiency, the pulse-forming network is usually operated at a high power level, and its output pulse is applied to the transmitting tube through a pulse transformer. On the other hand, the blocking oscillator is almost always used at a low power level and made to drive a hard-tube switch. There is reason to believe that blocking oscillators operating at high power levels would be practical for some applications.

Present beacon modulators are therefore divisible into two categories:

1. The "line type" modulator, consisting of a pulse-forming network and a gas-filled switch.

2. The "hard-tube" modulator, consisting of a blocking oscillator driving the grid of a hard-tube switch.

**12·7. The Line-type Modulator.**[1]—Since the line-type modulator is simply a pulse-forming circuit designed to drive the transmitting tube, the basic principles involved in its design have already been discussed above. The application of these principles will be clarified by an example. Suppose that a modulator is required for a beacon subject to random interrogation, that no coding is required, and that the modulator is to have the following characteristics:

Pulse power output = 1000 watts into 1000 ohms.
Pulse duration = 1 μsec.
Duty ratio = 0.1 per cent.
Power-supply voltage = 300 volts.

The random interrogation of the beacon makes resistance charging imperative. A duty ratio of 0.1 per cent can be achieved with a blanking gate of 1000 μsec, which may not be long enough to permit the use of resistance charging. We shall therefore compute the characteristics of a suitable network, and then check to insure that it will recharge in 1000 μsec. The circuit is shown in Fig. 12·5 (Sec. 12·4).

The efficiency of the output transformer must first be reckoned with. A well-designed transformer of the type needed here will have an efficiency of about 85 per cent; the effect of the losses will be to increase the primary current by about 10 per cent and to shorten the pulse in the load by about 5 per cent.

Although no adequate pulse specifications on the 2D21 gas-filled tube exist at present, it is the only tube available now which might be suitable for a circuit of this kind. Its suitability can be verified only by experiment.

During the pulse, the voltage drop across the switch tube is about 30 volts, regardless of the current. Therefore the pulse current, allowing for transformer efficiency, is

$$I_p = 1.1\,\frac{1000\text{ watts}}{(150\text{-}30)\text{ volts}} = 9.2\text{ amp.}$$

The 2D21 tube has been found to handle 10-amp pulses satisfactorily, so the choice of this tube still looks reasonable. The average current at the full duty ratio of 0.1 per cent is 9.2 ma, which is well within the 2D21 rating of 100 ma. Let $Z_p$ be the impedance presented by the primary of the transformer, and recall that the transformer reduces the power of the pulse by 10 per cent. Then

---

[1] Secs. 12·7 and 12·8 by J. J. G. McCue and P. A. de Paolo.

$$I_p^2 Z_p = 1.10 \cdot 1000,$$

whence

$$Z_p = 13.0 \qquad \text{ohms.}$$

Since the drop of 30 volts in the switch tube carrying 9.2 amp makes the tube have an effective impedance of 3.3 ohms, the pulse-forming network discharging into the transformer primary and switch tube in series will match its load if its characteristic impedance $Z_N$ is

$$Z_N = 13.0 + 3.3 = 16.3 \qquad \text{ohms.}$$

The inductance $L$ and capacitance $C$ of each section of the network can be computed from the relationships

$$Z_N = \sqrt{\frac{L}{C}}$$

and

$$\tau = 2n\sqrt{LC},$$

where $\tau$ is the pulse length and $n$ is the number of sections in the network. Choosing $n = 3$, which is reasonable, and letting $\tau = 1.05$ μsec to allow for pulse shortening by the transformer, we have

$$\begin{aligned} L &= 2.8\ \mu\text{h}, \\ C &= 0.010\ \mu\text{f}, \\ C_N &= 0.030\ \mu\text{f}. \end{aligned}$$

The transformer must transform the 1000-ohm load on the secondary to a primary impedance of 13.0 ohms; the required transformer stepup ratio is therefore $(1000/13)^{1/2}$, or 8.8 to 1.

The minimum value of $R_c$ which is required to quench the discharge in the switch tube can be computed from the supply voltage and the minimum current required to maintain the discharge. For the 2D21, this current is about 40 ma, hence, at a minimum,

$$R_c = \frac{E_B}{I_{\min}} = \frac{300}{0.040} = 7500 \qquad \text{ohms.}$$

If the time required to recharge the pulse-forming network through this resistor is taken as $4R_cC_N$, then the recharging time is $4 \cdot 7500 \cdot 0.030$, or 900 μsec. Allowing 100 μsec for recombination in the switch tube, we need a total of 1000 μsec between pulses. This is just the gate length required to limit the duty ratio to 0.1 per cent. Therefore we are operating on the fringes of feasibility; it would be well to place some inductance in series with $R_c$ to ensure quenching and hasten recharging.

The efficiency of the modulator can be estimated as follows: The pulse-forming network puts 150-volt pulses into its load, of which 20 per cent appears across the switch tube and 80 per cent across the transformer primary. The efficiency of the transformer is 85 per cent. Therefore the efficiency with which energy stored in the network is transferred to the 1000-ohm load is $0.80 \cdot 0.85$, or 68 per cent.[1] But only 50 per cent of the energy taken from the voltage source can be stored in the network if resistance charging is used; the rest is dissipated by the resistor $R_c$. This consideration reduces the efficiency to 34 per cent. A further loss is incurred through heating of the resistor during the recombination period. This loss will vary from tube to tube; experience indicates that it may amount, on the average, to about 30 per cent of the power used to charge the network. When this loss is taken into account, the efficiency is about 25 per cent. The load draws 1 watt at full duty ratio. The network circuit therefore draws about 4 watts from the high-voltage supply. The heater of the 2D21 requires another 4 watts. The over-all efficiency is therefore something like 12 per cent.

A more conservative design could be obtained if the power-supply voltage $E_B$ could be raised to 400 volts; for this case, the constants become

| | |
|---|---|
| Pulse current | 6.5 amp |
| Network impedance | 31 ohms |
| Inductance per section | 5.4 μh |
| Capacitance per section | 0.0056 μf |
| Charging resistor (minimum) | 10,000 ohms |
| Charging time (minimum) | 700 μsec |
| Transformer ratio | 1 to 6.2 |

This circuit has several advantages over the one designed for 300-volt operation. The pulse current in the 2D21 is smaller, satisfactory design of network and transformer is easier, and quenching can be ensured by making the charging resistor larger than the minimum value. The improvement in efficiency is trivial.

It would be misleading to leave the subject of line-type modulators without noting that the design of pulse-forming networks is an art as well as a science. The calculated values of the network constants serve simply as a point of departure; the final values must be reached empirically. Much helpful practical and theoretical information is to be found in *Pulse Generators*, Vol. 5, Chap. 6 of this series.

**12·8. Hard-tube Modulators.**—The hard-tube modulator consists of a pulse-generating circuit followed by an amplifier for raising the power level of the pulse. The pulse-generating circuit is almost always a blocking oscillator. The use of an amplifier has certain advantages: it can flatten the top of the pulse, and it makes the pulse shape relatively independent of changes in the load.

---

[1] This computation neglects energy losses in the capacitive part of the load.

Although the amplifier is of the ordinary Class C type, it is convenient to regard the amplifier tube merely as an electronic switch connected in series with the transmitting tube and a source of high voltage.[1] The switch has an appreciable inherent impedance; therefore, part of the supply voltage appears across the transmitting tube and the rest appears across the switch. Figure 12·9 shows a block diagram of a hard-tube modulator.

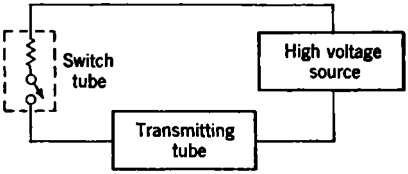


FIG. 12·9.—Block diagram of a hard-tube switching circuit.

Except for spontaneous frequency shifting, the problem of frequency stability is that of stabilizing the current pulses in the transmitting tube. There are two approaches to this problem: one is to prevent the supply voltage from varying with line voltage or with duty ratio, or from varying from pulse to pulse in a code group. The second is to use a switch that passes constant current, that is, one whose impedance increases as the voltage across it increases. The first approach is adopted in triode switching, the second in tetrode switching.

*Triode Switching.*—Figure 12·10 shows a triode-switching circuit. The triode is normally cut off by its grid bias. When the blocking oscillator driver (of which $T_1$ is a part) fires, a positive pulse is impressed on the grid of the triode, which becomes conducting and permits a current pulse to pass through $T_2$. A negative voltage pulse is therefore impressed on the cathode of the magnetron transmitter. The secondary of the transformer $T_2$ is a bifilar winding that permits the magnetron cathode to be heated by a grounded transformer $T_3$. The transformer $T_2$ is a stepup transformer which permits the use of a power supply, HV, the voltage of which is less than the operating voltage of the magnetron. The supply voltage should be kept as small as practical, since it must be

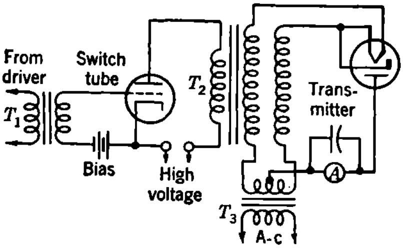


FIG. 12·10.—Triode switching and transformer coupling. Note bifilar transformer winding which permits heating of the transmitter cathode from a current source near ground potential.

---

[1] A discussion of hard-tube switches is given in *Pulse Generators*, Vol. 5, Chap. 3, Radiation Laboratory Series.

regulated to prevent it from varying with line voltage or with duty ratio. Since such regulation is difficult and expensive, triode switching is almost never used in beacon modulators. Except for very low voltage transmitting tubes, tetrode switching is much more convenient.

*Tetrode Switching.*—The difference between triode and tetrode switch tubes lies in their plate current–plate-voltage characteristics, of which typical examples are shown in Fig. 12·11.

Suppose that a triode switch is operating properly and then the high voltage increases. The current-voltage characteristic curve of a magnetron is such that a large change in current results from a small change in voltage. An appreciable part of the increase in supply voltage, therefore, will appear across the triode, moving its operating point from *A* to *B*. The current in the triode and in the magnetron will experience considerable increase; the result will be a change in frequency of the magnetron r-f output.

Suppose, on the other hand, that a tetrode is used as a switch tube. Almost all of the increase in supply voltage will appear across the switch tube, moving the operating point from *A′* to *B′*. It is clear that the current in the switch tube and in the magnetron will change very little; the transmitter will stay on frequency. As long as the operating point of the tetrode stays on the flat part of the characteristic, changes in supply voltage will have no appreciable effect on the magnetron current. The high voltage supply, therefore, need not be well regulated. It is sufficient to regulate the screen voltage, which is easy because the voltage is low and the current small. A modulator using this type of switching is sometimes called a "constant-current modulator."

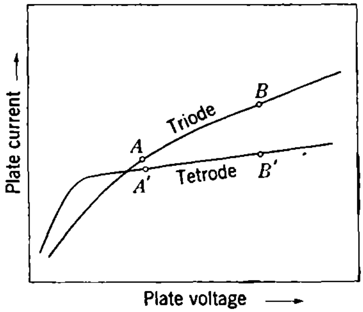


FIG. 12·11.—Triode and tetrode positive-grid characteristics.

An important aspect of operation on the flat part of the tetrode characteristic is that the current in the tube depends upon the grid drive. The pulse from the driver must therefore be shaped better than would be necessary with a tetrode operating on the steep part of the characteristic curve. The output current is approximately independent of variations in the high voltage, but it can be adjusted by adjusting the screen voltage or the drive on the control grid. The output power does vary somewhat with supply voltage but, since the current is nearly constant and most of the change in voltage appears across the switch tube, a 10 per cent change in supply voltage may cause as little as 5 per cent change in output power.

**12·9. Output Coupling Circuits.**[1]—The tetrode can be coupled to the magnetron through a pulse transformer, as in Fig. 12·10, or the direct-coupled arrangement of Fig. 12·12 can be used.[2] Here the switch connects the magnetron in series with a capacitor $C$ charged to an appropriate high voltage. The capacitor is made large enough so that it is only slightly discharged by each pulse; between pulses it is recharged by the power supply through the resistors $R_1$ and $R_2$. The inductance $L_c$ is made large enough so that the current in it does not build up to a very large value during the pulse; therefore, most of the pulse current goes through the magnetron. Furthermore, the current in $L_c$ tends to remain constant even after the switch tube is cut off; this results in the rapid discharge of the distributed capacitance $C_s$ associated with the circuit and thus the trailing edge of the pulse is steep. The value of $L_c$ must be made large enough to present a high impedance to the pulse but small enough to give a sufficiently steep trailing edge. The resistor $R_2$ is adjusted to damp critically the oscillations induced in $L_c$ and $C_s$.

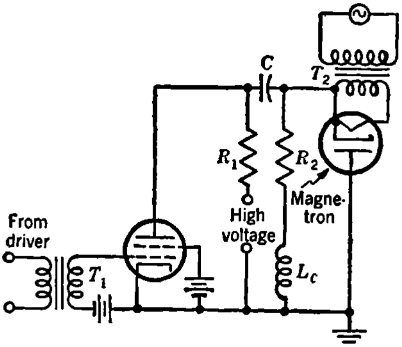


FIG. 12·12.—Tetrode switching with direct coupling. The pulse energy is stored in $C$, which recharges between pulses through $R_1$, $R_2$, and $L_c$. $T_2$ is a low-capacity transformer.

The secondary of transformer $T_2$, which supplies the heater current for the magnetron, must have a very low capacitance to ground because this capacitance is in parallel with the magnetron. It must also be insulated to withstand the voltage across the magnetron.

A pulse transformer may be desirable to permit the use of a low power-supply voltage, but it lowers the impedance into which the modulator works and therefore may lower the efficiency. Maximum power transfer, which is frequently more important than maximum efficiency, is obtained by matching the load impedance to the internal impedance of the switch-tube circuit. A constant-current modulator usually requires a transformer ratio near 1 to 1, and this ratio is therefore approximated with direct coupling.

When the modulator and transmitter are far apart, it is desirable to connect the modulator to a stepdown transformer that works into a low-impedance line, with the transmitter connected to the other end of the

---

[1] By J. J. G. McCue, P. A. de Paolo, K. R. More, and J. C. Reed.

[2] The design of output circuits is treated in considerable detail in *Pulse Generators*, Vol. 5, Chap. 2, of this series.

line through a stepup transformer. Here the modulator works into a high impedance, but the efficiency suffers nevertheless because a loss of 5 to 10 per cent is incurred in each of the two transformers. The choice of output circuit should, in general, be governed by the consideration that direct coupling is more efficient than transformer coupling and causes less distortion of the pulse.

*Triode Transmitter.*—In the foregoing paragraphs of this section it has been assumed that the transmitting tube is a magnetron. The same considerations apply to pulsing the plates of triode oscillators except that in this instance the need for a positive pulse makes transformer coupling mandatory unless a cathode-follower arrangement of the switch tube is used; also, the starting time of the oscillations may be longer than for magnetrons. If the triode oscillator is operating near its low-voltage limit, the starting time of the oscillations may be several tenths of a microsecond. This effect can be allowed for by making the drive pulse longer than the desired r-f pulse or by reducing the starting time through an increase in the pulse voltage.

*Tube Choice.*—The choice of a switch-tube is inseparably connected with the power-supply voltage and determines the choice of output circuit. The family of plate-current–plate-voltage curves for various control-grid and screen-grid potentials constitutes the crucial information about the switch tube. It is clearly desirable that the sharp bend in the characteristic curve occur at a low plate voltage and that the slope of the curve for higher plate voltages be small. It is also desirable that the ratio of plate current to control-grid current be high, in order that the control grid may be driven by a blocking oscillator of moderate size. The tube must pass the desired plate current for reasonable values of the voltages of the control grid and screen grid. Care should be taken to choose a tube with a plate well enough insulated to withstand the full voltage of the power supply. At present the number of tube types suitable for any given application is limited.

Table 12·1 lists some of the switch tubes, together with suitable drivers, that have been used successfully in beacon modulators.

TABLE 12·1.—HARD SWITCH TUBES AND DRIVERS

| Switch tube | Blocking-oscillator tube | Load current, amp | Load voltage, kv | Supply voltage, kv | Ratio of coupling |
|---|---|---|---|---|---|
| 3D21A | Two 6C4's | 1.0 | 2.5 | 2.5 | 1 to 1.5 |
| 715C | 807 | 3.0 | 4.5 | 6.7 | Direct |
| Two 715C's | 3E29 | 10 | 10.5 | 15 | Direct |

**12·10. The Blocking-oscillator Driver.**[1]—The current required by the control grid of the switch tube furnishes a guide to the type of tube required in the blocking-oscillator driver stage of the modulator, but the choice of this tube type cannot at present be based on computation.

On the basis of Table 12·1 and such other information of this kind as may be obtainable, an experimental modulator should be constructed, and a few trials made, to find a suitable blocking-oscillator tube. The output obtained from a given tube can be varied over rather wide limits by adjusting the plate voltage. By experimenting with different blocking-oscillator transformers and different line impedances, the designer can shape the pulse on the grid of the switch tube in such a way as to produce a satisfactory current pulse in the transmitter tube. Large leakage inductance in the blocking-oscillator transformer causes a pulse with a long rise time and a rising top. The rising top may compensate for a "droop" introduced in the output transformer, and a moderately long rise time may be desirable in order to eliminate a peak at the beginning of the current pulse or in order to eliminate spontaneous frequency-shifting. It is sometimes advisable, therefore, to put 10 to 250 $\mu$h in

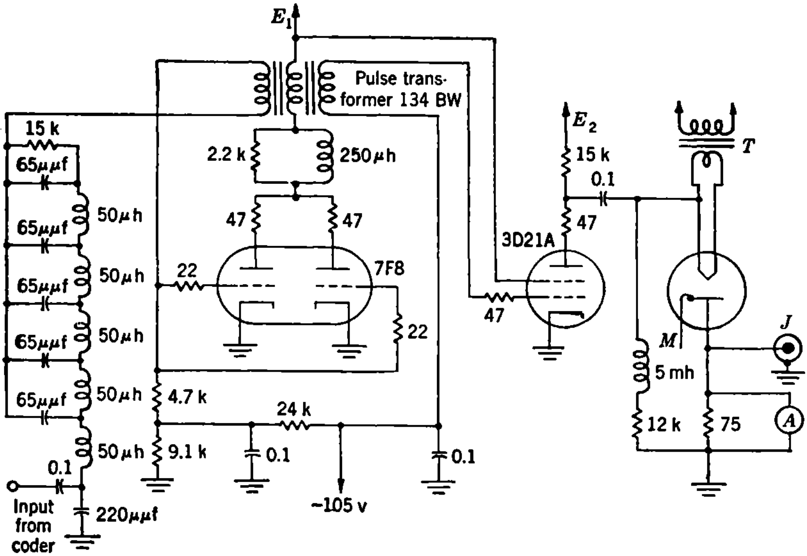


FIG. 12·13.—A typical hard-tube modulator, delivering about 2-kw pulses. 7F8—blocking-oscillator driver; 3D21A—tetrode switch tube; *T*—low-capacity transformer; *M*—magnetron transmitter (2J41); *J*—monitoring point; $E_1$—adjustable regulated supply, about 350 volts; $E_2$—3.5 kv, unregulated.

---

[1] Secs. 12·10 and 12·11 by J. J. G. McCue and P. A. de Paolo.

series with the transformer primary in order to increase the effective leakage inductance. The added inductance may be critically damped with a shunting resistance. If the blocking oscillator uses two tubes in parallel, resistances of about 20 ohms may be placed in series with the individual grids and plates to prevent parasitic oscillations and to equalize the powers contributed by the two tubes. Oscillations set up in the transformer can be damped out by placing resistors in parallel with one or more of the windings.

Too large a grid bias will prevent triggering of the blocking oscillator, but too small a bias will permit the oscillator to fire without being triggered. The bias should be set near the middle of the range which gives satisfactory operation. Since the pulse duration is slightly dependent on bias, precise adjustment of pulse duration can be made by carefully selecting the bias.

After the circuit constants have been chosen, the modulator should be tested to ensure that the modulator output does not vary appreciably with duty ratio, with the code employed, or with line voltage. A typical circuit of a complete hard-tube modulator is shown in Fig. 12·13.

*Efficiency.*—The efficiency of a tetrode modulator cannot be estimated easily. In fact, one of the chief advantages of the tetrode modulator is that its efficiency changes automatically when the supply voltage changes. Nevertheless, the factors influencing efficiency can be considered qualitatively and rough quantitative estimates can be made.

The high impedance of the hard-tube switch is its principal characteristic. When maximum transfer of power is more desirable than maximum efficiency, the load impedance should be matched to the tube impedance, to get an efficiency in the output circuit of 50 per cent. When high efficiency is desired, the load impedance should exceed the switch impedance by as much as possible. Direct coupling is clearly desirable. In a direct-coupled modulator, the efficiency of capacitor discharge into the load is likely to be near 70 per cent with present tubes. (Knowledge of the load impedance, the characteristic curves of the switch tube, and the impedance of the charging circuit permits the efficiency to be calculated in each instance.) The efficiency of charging the storage capacitor is hard to estimate, but it is much higher than the value of 50 per cent which characterizes resistance-charging in a line-type modulator because the storage capacitor loses only a small fraction of its charge during the pulse.

The foregoing considerations tend to favor the hard-tube modulator, as compared with the line-type modulator. Nevertheless, when the power expended in heating the cathodes of the switch tube and blocking oscillator, and the power dissipated by the switch-tube screen and the blocking-oscillator plate are taken into account, it is found that the hard-tube modulator has about the same efficiency as a line-type modulator that uses resistance charging.

**12·11. Comparison of Line-type and Hard-tube Modulators.**—The only general rule to guide the designer in his choice of modulator is that the choice must depend on a detailed analysis of the requirements the modulator must meet. The bases for the choice can be summarized as follows:

I. *Advantages of the Line-type Modulator*
    A. It is small and light
    B. It uses a simple circuit
    C. It operates from relatively low voltage
    D. It has relatively high efficiency when resistance charging can be avoided
II. *Disadvantages of the Line-type Modulator*
    A. In its simple form, it requires a long time between pulses
    B. It introduces a delay which may be as long as 1 μsec and may change when the switch tube is changed
    C. Its output power depends on supply voltage
    D. It is not very tolerant of changes in load impedance
    E. The top of the output pulse is not very flat
    F. In high-power circuits the pulse width is not easy to change
    G. It becomes complex for multiple-pulse codes
III. *Advantages of the Hard-tube Modulator*
    A. It will function at high and variable repetition rates
    B. It introduces only a small delay
    C. Its output power is nearly independent of supply voltage over a considerable range, provided the proper part of the tetrode characteristic is used
    D. It is flexible as to load impedance
    E. The top of the output pulse can be made very nearly flat
    F. The pulse width is easily changed
    G. It is well suited to multiple-pulse codes
IV. *Disadvantages of the Hard-tube Modulator*
    A. It is bulkier and heavier than a line-type modulator with the same output power
    B. It uses a relatively complex circuit
    C. It requires a high supply voltage and one or more auxiliary voltages

# CHAPTER 13

## BEACON TRANSMITTERS: MAGNETRONS

BY R. DICKINSON, K. R. MORE, P. A. DE PAOLO, AND J. C. REED

**13·1. General Considerations.**[1]—The beacon transmitters to be discussed below cover the frequency range from about 100 to 30,000 Mc/sec. The types of tube and circuit used depend on the frequency. For frequencies from 100 to about 3000 Mc/sec, specially designed triodes and suitable external circuits, can be used. Magnetrons are used for frequencies from about 2000 Mc/sec to 30,000 Mc/sec or more. When cavity magnetrons are used, the oscillating circuits are contained within the tube. Both magnetrons and triodes are used in the 3000-Mc/sec region. The choice is usually determined by availability, power requirements, and restrictions of weight and size on the beacon.

Requirements for transmitted pulse power are generally lower for a beacon system than for a radar system because only one-way transmission is used. The pulse-power output required varies with the frequency used and with the type of interrogator. The ranges of pulse power that have been used are 3 to 10,000 watts at 100 to 300 Mc/sec, 15 to 5000 watts at 500 to 1200 Mc/sec, 50 watts to 20 kw at 3000 Mc/sec, and 300 watts to 40 kw at 10,000 Mc/sec. The wide variations are due to variations in the receiver sensitivity, antenna gains, and displays of the interrogators, as well as in the required range. Large ground or ship radars with high antenna gains and sensitive receivers can pick up replies from relatively low-power distant beacons. On the other hand, airborne radars with small antennas and perhaps less sensitive receivers require a larger power output from beacons if the replies are to be received properly.

While the pulse power for a beacon transmitter is relatively low, the average power is greater in relation to the pulse power than for a radar set because of the high interrogation rates possible and the use of a multiple-pulse reply code. Duty ratios as high as 1 per cent or more are used, compared with the usual radar duty ratios of 0.1 per cent or less.

The requirements for frequency stability are more stringent for beacons than they are for radars because beacons usually must use fixed reply frequencies. Since many interrogator receivers have bandwidths of 2 Mc/sec or less, beacon transmitters must often be on the correct fre-

---

[1] Secs. 13·1 to 13·10 by K. R. More and P. A. de Paolo.

quency to within ±1 Mc/sec. It is often difficult to attain this degree of stability in view of the frequencies used and the wide range of interrogation rates possible. Change of frequency with duty ratio, power supply voltage, r-f load, and ambient temperature must be kept to a minimum.

The use of a fixed reply frequency also makes it necessary that the beacon transmitter be tunable, to some extent. Tuning sometimes is done over a range only great enough to tune any tube of a given type to the desired frequency; when more than one fixed frequency must be available, however, tuning over a fairly wide range is required.

The important general properties of magnetrons[1] and various special beacon requirements and applications are discussed briefly in this chapter. Discussions of triode tubes and oscillator circuits used in beacons follow in Chap. 14.

**13·2. Performance Charts and Rieke Diagrams.**—The performance of a magnetron may be described in terms of seven variables: magnetic field, applied voltage, input current, wavelength, power output, load resistance, and load reactance. The last two are frequently referred to as "the load," and are determined by the r-f line and load coupled to the output of the tube.

Data on magnetron performance are conveniently presented graphically by means of performance charts and Rieke diagrams. Performance charts are obtained by fixing the load, varying the magnetic field and voltage, and measuring the resulting values of input current, wavelength, and power output. The data can then be presented graphically as performance charts by plotting the pulse voltage as ordinate against the pulse current as abscissa. The magnetic field, power output, efficiency, and wavelength are marked for each point for which data are taken. Contours of constant magnetic field, constant power, constant efficiency, and constant wavelength are drawn. A typical performance chart is given in Fig. 13·1. The principal region of poor performance is indicated on the chart. The regions of poor performance frequently are those in which the tube shows more than one frequency of oscillation.

Performance charts are convenient for choosing the most suitable operating values of voltage, current, and magnetic field for a given magnetron in order to obtain a desired power output from a given magnetron. They are also useful in making the choice of magnetron to be used for a given purpose, considering the power output desired and the pulse voltage and current that can be made available to drive the tube.

The data for Rieke diagrams are taken by varying the load while keeping the magnetic field and input current constant. The wavelength, power output, voltage standing-wave ratio in the line, and the position

---

[1] The subject of magnetrons is taken up in detail in *Microwave Magnetrons*, Vol. 6, Radiation Laboratory Series.

of a minimum of the standing wave as referred to the output coupling are measured. The values of the phase and amplitude of the standing wave represent the load, since the standing wave is determined by the impedance that the line presents to the magnetron. The data are plotted on circle diagrams as illustrated in Fig. 13·2. The power standing-wave ratio ($r^2$) is plotted as the radial coordinate. The angular coordinate is computed from the position of the standing-wave minimum measured from an arbitrary reference point. It is expressed in degrees, with one-half wavelength equal to 360°. The angular coordinate increases as the minimum moves away from the load. Contours of constant frequency, constant power, constant voltage, and constant efficiency may be drawn.

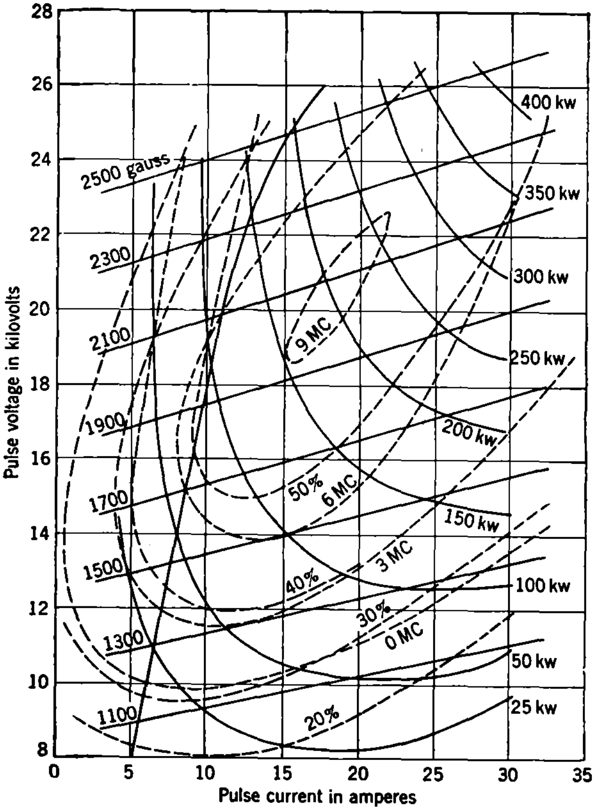


FIG. 13·1.—A typical magnetron performance chart. The family of parallel straight lines indicates the operating points as a function of magnetic field. Contours joining points of equal efficiency, power output, and frequency are drawn in. The principal region of unsatisfactory operation, as evidenced by a poor spectrum, is the area to the left of the steep diagonal line which starts at 8 kv, 5 amp. Operation is with a matched load.

For any system, the choice of the operating point is a compromise. The region of maximum power output frequently occurs near the region of frequency instability and high standing-wave ratio. In such a case a point of lower power and reasonable frequency stability must be chosen.

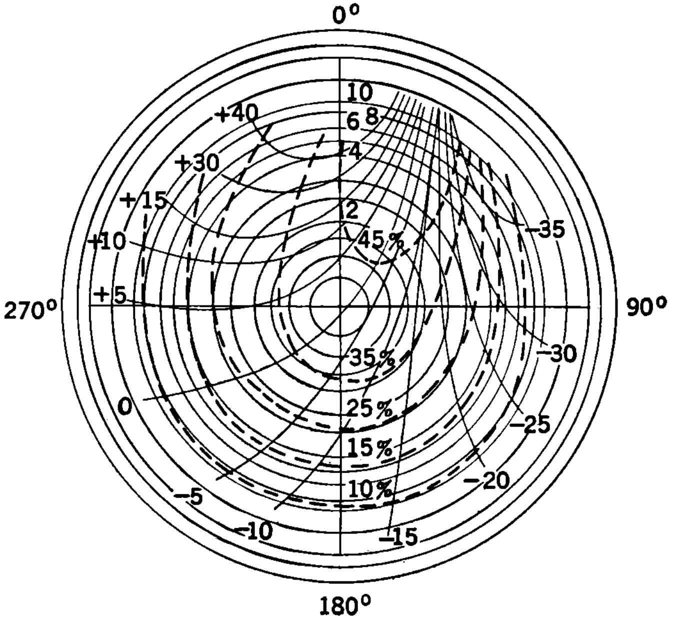


Fig. 13·2.—Rieke diagram of a 2J41 magnetron. The radial coordinate is the reflection coefficient of the load (actually values of power standing-wave-ratio are shown). The angular coordinate is the position of a minimum in the standing-wave pattern with respect to the magnetron output coupling. The full lines are contours of frequency shift in Mc/sec, as referred to the matched-load frequency. Dotted lines show contours of constant efficiency. The input current, pulse-repetition frequency, and pulse length are fixed.

**13·3. Frequency Stability.**—The frequency at which a given magnetron oscillates depends upon the operating conditions and specifically upon the load, magnetic field, and tube current. Because of the thermal expansion of the resonant cavities of the anode, the frequency also depends upon the ambient temperature and upon the average power dissipated in the tube. For a 10-cm magnetron with a copper anode the frequency decreases about 0.06 Mc/sec per °C rise in the temperature of the block, and about 0.18 Mc/sec per °C for a 3-cm tube. Methods of increasing the frequency stability to the degree required for beacon operation will be discussed below.

*"Pulling Figure" and "Pushing Figure."*—The "pulling figure" of a magnetron is defined as the maximum change in frequency which occurs

when the load is varied in such a way as to maintain a voltage standing-wave ratio of 1.5 while varying the phase over a range of 360°. Pulling figures of available 3-cm magnetrons range from 6 to 15 Mc/sec and from 4 to 10 Mc/sec for 10-cm magnetrons.

The frequency shift, or pulling, as the load changes, may be very serious in system operations because r-f loads are subject to accidental variations. Pulling, however, can be used as a means of tuning over a narrow range by controlled changes in the load, as will be discussed.

The "pushing figure" is defined as the frequency change for a current change of 1 amp when the magnetic field and load are held constant. This is of importance in determining the degree of regulation of the modulator and power supply, with respect to changes in input voltage and duty ratio, needed to obtain a given desired frequency stability.

Pushing figures vary not only from one type of tube to another, but also with the operating point for any given tube. In choosing the operating point, the pushing figure and the probable current variation to be expected from the power supply must be kept in mind. A typical pushing figure for a typical 10-cm 12-kv magnetron in its normal operating region is 3 Mc/sec per amp. With many types of tubes operating as recommended, pushing figures are negligible.

*Mode Changes.*—The types of frequency instability discussed above give rise to shifts in frequency amounting to a fraction of a per cent. In addition, there is the possibility that the magnetron may change its frequency—sometimes erratically—by 10 per cent or more *by changing its mode of oscillation.* The mode-selection properties of magnetrons vary with the type of the tube and also depend upon the operating conditions. Mode changes may set both upper and lower limits to the current at which operation is satisfactory, but these limits often may be altered by changing the internal resistance of the modulator and the rate at which its output voltage rises at the beginning of the pulse.

*Spectrum.*—The spectrum of a magnetron gives a measure of the frequency distribution of the emitted energy. It can be plotted graphically as intensity versus frequency. The theoretical distribution, which has the form $\sin^2 x/x^2$ in power for a rectangular pulse, is shown in Fig. 4·1, Sec. 4·11. The distance between the first minima in Mc/sec is $2/t$, where $t$ is the pulse duration in microseconds. The spectrum of a good magnetron approaches that theoretically expected, though the relative intensities of the secondary maxima frequently exceed the theoretical values.

Certain conditions that lead to a poor spectrum should be avoided. One is operation near regions in which unwanted frequencies occur. Some defective tubes show additional regions of bad spectrum for certain values of voltage and current. Poor pulse shape can give rise to a broadened spectrum if the tube is operating in a region in which the frequency changes rapidly with current.

**13·4. Magnetron Input Requirements.**—It has been mentioned previously that mode selection is influenced by the rate at which the voltage applied to the magnetron rises at the beginning of the pulse; in addition, the tendency of the magnetron to spark may be enhanced if the voltage rises too rapidly. The sparking may shorten the life of the tube or cause the frequency of the magnetron to drift during the life of the tube through the deposition of material on parts of the oscillator structure.

Both sparking and mode-changing are connected with the facts that a finite time is required for oscillations to build up in the magnetron, and that the buildup can continue only as long as the applied voltage remains within a rather narrow range. If oscillations do not build up completely and thus cause the magnetron to draw current sufficient to load the modulator, the voltage applied to the magnetron may reach abnormally high values, particularly if the modulator is of a type which normally operates with a large internal voltage drop. At the higher voltage, the magnetron may start oscillating in some undesired mode or it may spark.

Trouble of this kind can be minimized by the use of inductance to limit the rapid rise in the current, and by the use of a capacitor that is charged from the pulse voltage supply through a resistor in order to reduce the rate of rise of the voltage.[1]

Sparking difficuities are less likely to occur with hard-tube modulators because of their different regulating properties. The voltage rises only a small amount above the operating value during the starting interval.

Not all magnetrons have the same starting interval. Some types begin to oscillate at low power before the voltage reaches the operating value, with the power rising rapidly as the voltage increases. The starting interval varies as the operating conditions are changed, even for a given tube.

Even if sparking does not occur, too rapid a rise in the voltage pulse may result in a gradual drift of the frequency of oscillation, usually to lower values, accompanied by a tube life too short for satisfactory performance. This can be avoided by reducing the rate of rise of the voltage pulse, particularly near the top of the pulse. These phenomena are apparently associated with rapid deterioration of the cathode that may result from excessive evaporation of the cathode coating. Good cathodes do not show this trouble to any extent, although many inferior cathodes have been used in production tubes in the past.

---

[1] These magnetron-modulator interactions are taken up in more detail in *Microwave Magnetrons*, Vol. 6, Chap. 8, Radiation Laboratory Series.

## SPECIAL BEACON MAGNETRON REQUIREMENTS

We have seen that magnetrons for beacon service must satisfy certain requirements not encountered in ordinary radar sets. In most cases the power output required is less than that required of magnetrons for radar use, because beacons use one-way transmission. For airborne beacons, weight is also an important factor. Beacon transmitters frequently work at high duty ratios because beacons use multiple-pulse reply codes and must respond to high interrogation rates. The common beacon practice of having a standard fixed reply frequency places more stringent requirements on tunability and frequency stability than for radar sets.

**13·5. Coded Operation.**—When a magnetron is used as a beacon transmitter, it may be called upon to transmit a series of r-f pulses which are relatively closely spaced. The commonly used code spacings vary from about 10 to 45 μsec. The use of multiple-pulse codes brings up two problems in connection with magnetron performance. First is the duty ratio, which may be as high as 0.6 per cent for a six-pulse code with 0.5-μsec pulses and an interrogation rate of 2000 cps. The design of the magnetron must take into account the duty ratio and the pulse power at which the tube will be operated. Also, the changes in duty ratio resulting from changes in interrogation rate, as mentioned above, may result in a change in the frequency of the transmitted r-f power. When this is excessive, steps must be taken to reduce the shift.

The second problem arising because of coded operation is that of the equality of the pulse power emitted during the different pulses of the code. Experiments with present beacon magnetrons have shown that the pulse power outputs are equal, for the pulse spacings used, provided the pulse current drawn by the tube is not too great. When the current is increased beyond a certain value, the power output of the last pulse of a six-pulse code may be 10 to 20 per cent below that of the first pulse. In view of the relatively low power output required of a beacon transmitter, it is usually possible to operate at currents low enough for satisfactory performance.

**13·6. Tuning.**—The use of fixed-frequency transmission in beacon services makes it necessary to provide some means of tuning the magnetron to the desired frequency. Two methods of tuning are available. One is the use of a tunable magnetron. The other is the use of controlled variations of the magnetron load to pull the frequency to the desired value.

Whenever frequency stabilization of the magnetron is desirable, which is almost invariably the case, tunable magnetrons are preferable to fixed-tuned magnetrons. The attainable stabilization is better if the stabilizing device does not have to pull the magnetron frequency as well as stabilize it. In the past fixed-tuned magnetrons have been used in beacons largely because tunable tubes were not available.

*Tunable Magnetrons.*—One method used for tuning magnetrons involves changes in the internal resonant circuits of the magnetron by a mechanism that can be controlled externally. Another method involves the use of a tunable cavity coupled to the magnetron cavity, the two cavities forming a common evacuated system. A third method involves the use of electronic tuning. Some of the mechanisms demanded by these methods, however, add appreciably to the size and weight of the tube, which may be a drawback in lightweight beacon systems.

*Mechanical Tuning.*—The tuning range available with mechanical tuning[1] is limited by the drop in power output at the ends of the tuning range, and in some cases by mode trouble. In general, tunable magnetrons now available have a usable tuning range of 5 to 10 per cent of the center frequency. With this amount of tuning, the power output at the ends of the tuning range will be 10 to 20 per cent below the maximum power output near the center of the range. Some tubes give a tuning range up to 12 per cent, with a drop of power of about 50 per cent. A tuning range of 1 to 2 per cent is adequate for fixed-frequency beacon operation. Such a tuning range is readily achieved by a simple mechanism which introduces capacity loading of the resonant cavities in the end space.

*Electronic Tuning.*—As in the case of mechanical tuning, electronic tuning[2] may be applied to the resonant circuits of the magnetron or to an iris-coupled cavity which forms part of the magnetron vacuum chamber. Either a controlled beam of electrons is passed axially along one of the slots of the magnetron or the beam is passed along the axis of the cavity.

At present, electronic tuning is in an experimental stage. No tubes that use it are yet available. Should electronically tuned magnetrons become available for beacon use, very accurate automatic frequency control will become readily possible.

**13·7. Fixed-tuned Magnetrons.**—Methods of tuning fixed-tuned magnetrons depend on the changes of the operating frequency of a magnetron with changes in the impedance of the output load, as discussed in Sec. 13·3. Controlled changes in load impedance thus can be used to tune the magnetron. Frequency variations of a fraction of 1 per cent are possible by this method. Although this is not a wide tuning range, it is sufficient for many beacon services if the tubes chosen have matched-load frequencies close to the desired beacon frequency.

---

[1] Mechanical tuning is discussed at length in *Microwave Magnetrons*, Vol. 6, Chap. 14, Radiation Laboratory Series.

[2] Electronic tuning is discussed in detail in *Microwave Magnetrons*, Vol. 6, Chap. 15, Radiation Laboratory Series.

*Stub Tuners.*—A single-stub tuner correctly placed along the transmission line provides a convenient means of varying the load impedance. With this method, tuning is accomplished by one adjustment. The power-output variations are not excessive. Double-stub or double-slug tuners are not convenient in actual use because they require two adjustments, each of which influences both the frequency and power output. The use of the single-stub tuner is adequate when frequency stabilization is not required.

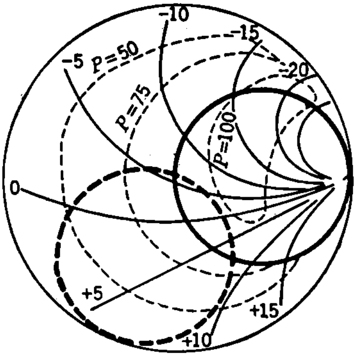


FIG. 13·3.—Rieke diagram showing the operation of a tuning stub. The frequency contours are labeled with respect to the beacon frequency. The matched-load frequency of the magnetron is about −6 Mc/sec referred to the beacon frequency. The dotted lines are power-output contours. The heavy circles represent the effect of tuning the stub; the full heavy circle is for a properly located stub, and the heavy dotted circle for an improperly located stub. When the stub is properly located at the frequency sink, maximum tuning effect is obtained. The operating point is the intersection of the heavy circle with the contour of zero frequency shift.

The impedance presented to the magnetron by a properly located variable stub and a matched line is represented by points on the heavy solid circle in the Rieke diagram shown in Fig. 13·3. Tuning the stub moves the operating point around the circle, with a corresponding frequency change. Stable operation and smooth tuning are obtained for points far away from the region of greatest frequency sensitivity, the frequency "sink," which is the point toward which the frequency contours converge.

If the stub is not properly located the impedance is represented by points on a circle like the heavy dotted one. In such a case the range of tuning for stable operation is less than for a properly located stub. The location is not critical, however, and a suitably chosen average position is possible for tube types that show a reasonable degree of uniformity. The 10-cm tubes now available often exhibit sufficient uniformity; 3-cm tubes do not.

The correct position for the stub can be determined by constructing a Rieke diagram for the magnetron. The point toward which the frequency contours converge corresponds to a minimum in the standing-wave pattern at a particular position along the line. The stub should be connected at such a point. For best operation, a position as close to the magnetron as possible should be chosen. A second method is to make cold impedance measurements on the magnetron. The stub should be placed at a position of a standing-wave minimum for wavelengths well off resonance. For tubes with nonuniform characteristics, the correct position must be determined for each tube, either by the user or by the manufacturer.

*Cavity Tuners.*—A tunable cavity correctly placed along the transmission line provides another method for varying the load impedance. The use of a cavity having a loaded $Q$ greater than that of the magnetron also provides a considerable degree of frequency stabilization. A more complete discussion of the stabilizing tuner is given in Sec. 13·10.

**13·8. Frequency Modulation.**—In some cases it is desirable to introduce frequency modulation of the beacon reply frequency. This has been done when the frequency stability was inadequate to insure reception, and stabilization techniques were not yet developed. It may still be desirable as a coding parameter, as described in Chap. 5. However, frequency modulation is undesirable as a substitute for adequate stability because of the display losses involved. It is no longer necessary technically since adequate stabilization techniques are now available.

The frequency of either fixed-tuned or tunable magnetrons may be swept over a narrow band by the addition of a variable reactance in the line. This can be done in principle, for example, by driving the plunger of a stub tuner with a reciprocating motion. The use of a rotating resonant ring[1] in the transmission line is more satisfactory and can be done quite simply in the case of a waveguide system by placing the ring across the guide. For a coaxial transmission line, a short waveguide section containing the resonant ring can be iris-coupled to the line, as a stub. This waveguide stub should be located in the place that is best for the single-stub tuner.

With present designs it is not convenient to sweep the frequency of a tunable magnetron of the mechanical type by periodic driving of the tuning mechanism. It is, of course, quite easy to sweep the frequency of an electronically tuned tube by means of the electronic controls.

**13·9. Long-line Effect.**—In many beacon installations the line from the transmitting tube to the antenna necessarily has considerable length. The effects of mismatches of the line and antenna on the frequency of oscillation of the magnetron may be serious when the transmission line is long.[2]

When a magnetron is used with a long transmission line the tuning may become irregular, or even discontinuous if the mismatch is serious. Such irregular tuning is illustrated in Fig. 13·4 where the frequency of oscillation under operating conditions is plotted against the frequency which would be obtained for the corresponding tuning setting if the load

---

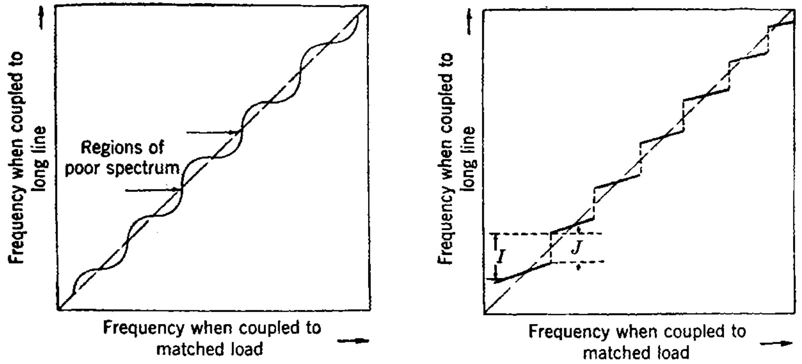
[1] Details of the theory and design of resonant rings are given in *Waveguide Handbook*, Vol. 10, Chap. 6, Radiation Laboratory Series.

[2] Discussed further in *Microwave Magnetrons*, Vol. 6, Chap. 7 of this series.

were perfectly matched. The tuning in Fig. 13·4*b* is discontinuous. In Fig. 13·4*a* it is continuous, but there are regions of very rapid change of frequency and poor spectrum. The effect of an unmatched line on tunable magnetrons is shown as a function of line length, degree of mismatch, and pulling figure in Fig. 13·5. From this figure the effect of any given mismatch may be determined. The interval $I$ between breaks in the tuning curve is given by the equation

$$I = \frac{150\lambda}{L\lambda_g}, \tag{1}$$

where $I$ is in megacycles per second, $\lambda$ and $\lambda_g$ are, respectively, the free-space and transmission-line wavelength, and $L$ the line length in meters.



(*a*) (*b*)

FIG. 13·4.—The long-line effect. The tuning of a magnetron coupled to a long line not perfectly matched may be irregular, as in (*a*), or even discontinuous, as in (*b*). (*a*) Tuning curve when magnetron is coupled to a long line with a slight degree of mismatch. (*b*) Tuning curve when a magnetron is coupled to a long line with a greater degree of mismatch. The degree of mismatch necessary to produce discontinuous tuning and the fraction $J/I$ of the tuning range which is missing are given in Fig. 13·5. If the tuning is discontinuous, a phase-shifter can be used to assure that the operating frequency may be obtained.

If there are no actual breaks in the tuning curve, as in Fig. 13·4, $I$ gives the extent of a tuning cycle. It is clearly preferable to operate to the left of the curve for $W = 0$, so that tuning will be continuous. When a long line is used, it is advisable to incorporate in it a phase shifter, so that the effective line length may be varied. If the desired frequency of operation falls in a break in the tuning curve or in a region of rapid change when no breaks occur, a change of the line length that need be no more than a fraction of a half wavelength will move the point of operation to the flat part of the curve. When the system operates on the flat part of the curve, the long line, acting in somewhat the same way as a stabilizer, prevents frequency changes. This type of stabilization is not very desirable because of its relative unmanageability.

One of the serious aspects of the long-line effect is its variation with temperature. The thermal expansion of a long transmission line with the normal temperature changes encountered in most climates is often enough to change the effective line length an appreciable fraction of a wavelength. This expansion may pull the magnetron far off its assigned frequency.

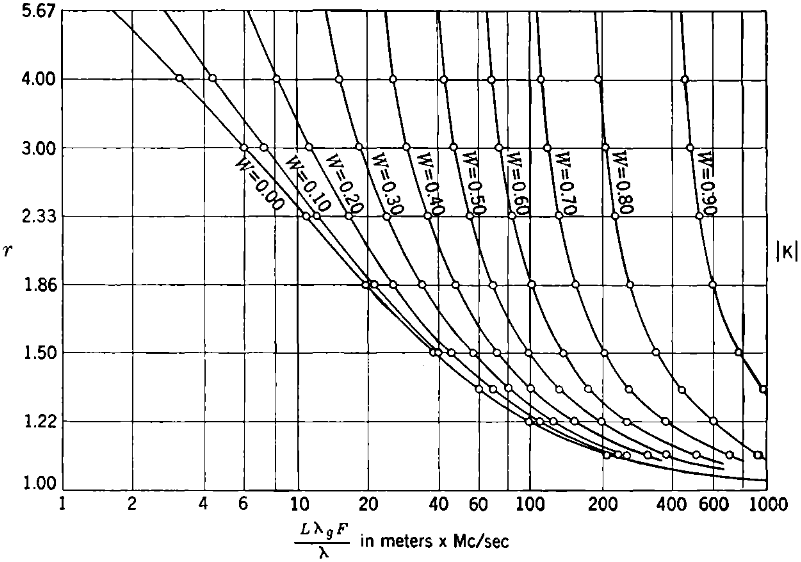


FIG. 13·5.—The magnitude of the long-line effect. The magnitude of the reflection coefficient of the load $|K|$ and the voltage standing-wave ratio $r$ are plotted against the function $\frac{L\lambda_g F}{\lambda}$. Curves are shown for various values of the degree of discontinuous tuning $W$. $L$ is the line length in meters; $\lambda_g$ is the transmission-line wavelength, $\lambda$ the free-space wavelength in the same units; $F$ is the pulling figure of the magnetron in Mc/sec. $W$ is defined as the ratio $J/I$, where $J$ is the width of the break in the tuning curve in Mc/sec, and $I = 150\ \lambda/L\lambda_g$ is the interval between breaks in the tuning curve (see Fig. 13·4) in Mc/sec. The mismatch is assumed to occur at the end of the transmission line.

There is no cure for this difficulty. It must be avoided by careful matching of the r-f load, and, if necessary, by controlled phase-shifting as the temperature changes.

**13·10. Magnetron Stabilization.**—Although it is nearly always necessary to keep the transmitted frequencies of beacons as close as possible to their assigned values, it is difficult to achieve the stability desired when a magnetron is used as a transmitting tube, for several reasons. A change of duty ratio causes a change in temperature of the magnetron and, therefore, a change in physical dimensions. Shifts in ambient temperature also cause changes in magnetron temperature and changes in the length of the line between the magnetron and the antenna. Poor regulation in the transmitter power supply and aging of the modulator tubes cause variations in magnetron pulse current, and, therefore, in frequency.

*Temperature.*—When a magnetron is delivering 100 watts of average r-f power with an efficiency of 20 to 40 per cent, about 150 to 400 watts of power are dissipated in the magnetron. At zero duty ratio this power is not present, which means that as the duty ratio varies large shifts of temperature and hence of frequency can occur. In a 3-cm magnetron, a change of 50°C in the temperature of the copper anode results in a frequency shift of about 9 Mc/sec; in a 10-cm magnetron about one-third this shift occurs. Thermostating is used to minimize temperature changes. The magnetron is kept at some temperature above the maximum operating temperature it would attain if no external heating were used. A cartridge-type heater is often used to keep the magnetron hot during periods of low duty ratio, and a blower to prevent excessive rises in temperature as the beacon is interrogated. Control within 5°C is adequate at 3 cm.

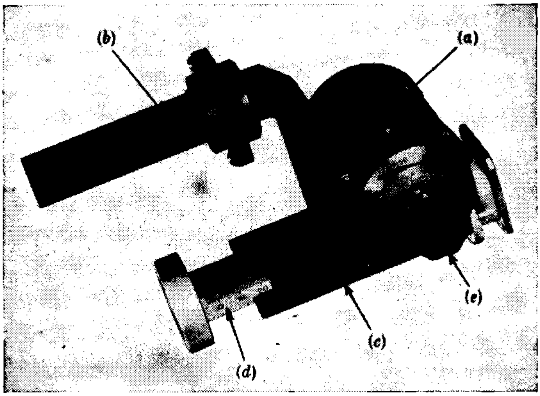


FIG. 13·6.—A 3-cm cavity stabilizing tuner. The stabilizing cavity (*a*) and associated matched load (*b*) are mounted on a sliding section (*c*) whose position is read from a scale on the fixed section (*d*). The 3-cm 2J48 magnetrons with which this tuner is used require individual adjustment of the slider setting. The tuning adjustment (*e*) tunes the cavity and the magnetron.

*The Stabilizing Tuner.*—It is possible to reduce the effect of all three general causes of magnetron frequency instability by coupling a tunable resonant cavity to the magnetron output. This cavity has the dual purpose of tuning the magnetron and stabilizing its frequency. The cavity, together with an associated dummy load, is called a stabilizing tuner[1] and is equivalent to a parallel-tuned circuit and resistance connected in series across the r-f line. When a stabilizing tuner is coupled at the correct distance from the magnetron, it exerts an appreciable frequency-stabilizing effect. The ratio of frequency change without a stabilizing cavity to frequency change with a stabilizing cavity is called the stabilization factor. Stabilizing tuners with stabilization factors from 3 to 10 or more, depending upon the tube used, have been built. It is necessary that the cavity used in the tuner be compensated for temperature changes; otherwise it will contribute to temperature-induced frequency instability by pulling the magnetron off frequency.

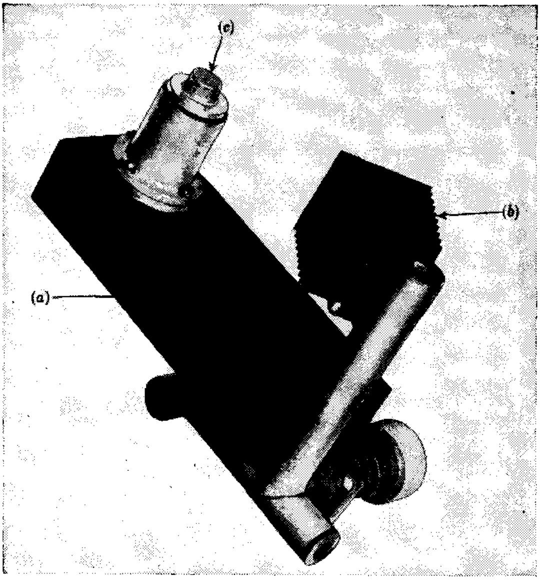


FIG. 13·7.—A 10-cm stabilizing tuner. In contrast to the 3-cm tuner, there is no sliding arm. The waveguide stabilizing cavity is fixed in position, since the 2J59 magnetrons used with this tuner have uniform characteristics. (a) Waveguide cavity. (b) Matched load. (c) Tuning adjustment.

Photographs of 3-cm and 10-cm stabilizing tuners that have been used in beacons are shown in Figs. 13·6 and 13·7, respectively. It will be noted that the cavity of the 3-cm tuner is mounted on a side arm that is adjustable by means of the sliding section, but that in the 10-cm tuner the position of the waveguide cavity is fixed. In either case, the cavity should be coupled to the line at a minimum of the standing-wave pattern, as in the case of the single-stub tuner. Fixed positions are possible for tube types

---

[1] The theory and design of stabilizing tuners is discussed in *Microwave Magnetrons*, Vol. 6, Chap. 16, Radiation Laboratory Series.

that show a sufficient degree of uniformity, as is the case with the 10-cm tubes (2J59) used with the stabilizer shown in Fig. 13·7. The 3-cm tubes (2J48) used with the stabilizer shown in Fig. 13·6 do not show sufficient uniformity to permit the use of fixed positions. The correct setting of the sliding section is stamped on each tube by the manufacturer.

Some data obtained with a thermostated tube and the stabilizing tuner shown in Fig. 13·6, are of interest. A change in duty ratio from 0.06 to 0.24 per cent causes a frequency shift of about 1.75 Mc/sec. Normally the beacon transmitter would be tuned at a duty ratio midway between the minimum and maximum limits, so that the shift in frequency would be centered about the correct frequency. Figure 13·8 summarizes the performance.

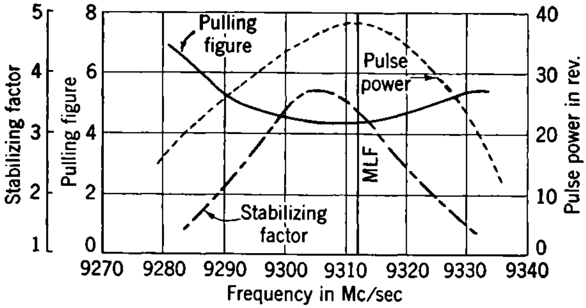


FIG. 13·8.—The stabilizing factor, pulling figure, and pulse power output of a 2J48 3-cm magnetron with a stabilizing tuner, as a function of frequency. The matched-load frequency (MLF) is 9312 Mc/sec.

## AUTOMATIC FREQUENCY CONTROL, AFC

BY R. DICKINSON, K. R. MORE, P. A. DE PAOLO, AND J. C. REED

The requirement of microwave beacons for frequency stability makes some kind of automatic control of transmitter frequency necessary if a very high degree of stability is to be attained under all conditions of operation. Because some radar receivers have bandwidths less than 2 Mc/sec, the transmitting frequency of a beacon working with such sets ought to be held to within ±1 Mc/sec of the assigned frequency under all operating conditions. The stabilizing tuner described in Sec. 13·10 is not quite adequate for this degree of stability. An AFC system can control the frequency to ±0.5 Mc/sec or better at 3 cm. Under heavy transient loads, the beacon-transmitter frequency may shift as much as 2 Mc/sec when using no AFC; under the same conditions, if AFC is used, the frequency shift will be less than 0.5 Mc/sec.

The systems to be described achieve frequency control by means of a servomechanism operated by an error voltage developed when the magnetron drifts away from the desired frequency. An AFC system consists of a discriminator that produces an error voltage when the transmitter is off frequency, an amplifier, and a servomechanism that operates from the information given by the amplifier. The servomechanism drives a tuner that brings the magnetron back to the desired frequency. A servomechanism is used rather than an electrical control, because available magnetrons cannot be tuned suitably by electrical means.

The methods for automatic control of frequency of beacon transmitters differ from those used with c-w oscillators for several reasons. First, because mechanical tuning is required, the time constant of the control system is necessarily increased. Second, the system must operate with pulses. If crystal detectors are used, the danger of crystal damage limits the pulse power which can be applied to the crystal. More important, the r-f information is no longer continuous, but consists of an amplitude-modulated series of pulses. Finally, the pulse-repetition frequency of beacons is not constant. The amount of information available is variable, and may be zero for long periods of time. All these factors must be taken into account in the design.

**13·11. AFC Discriminators.**—Discriminators for AFC of beacon transmitters at microwave frequencies differ from those customary at lower frequencies because oscillators in this region are not stable enough to be relied upon as frequency standards. The most accurate systems at the present time use a resonant cavity as a reference standard and approach the degree of stability that can be expected from the cavity.

The error voltage obtained from the discriminator must show when the magnetron is off frequency and in which direction (see Fig. 13·9). A simple high-$Q$ transmission cavity with a crystal-rectifier output is not a suitable discriminator. It tells when the oscillator is off frequency, but not in which direction. The systems to be described below each use a reference cavity, but additional means are provided to supply the necessary error information.

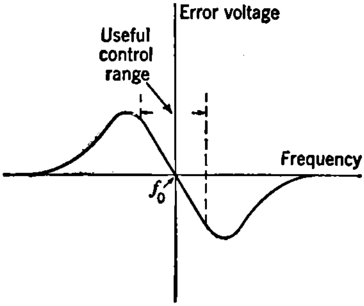


FIG. 13·9.—Discriminator characteristic. The discriminator error voltage changes sign as the frequency passes through $f_0$ at resonance.

*Modulated-cavity Single-crystal Discriminator.*—In the modulated-cavity discriminator, the frequency of a reference cavity is varied by a diaphragm driven like that in an earphone. The cavity is modulated over the range of frequency over which the transmitter is to be controlled by applying a low-frequency (60 to 400 cps) voltage to the coil that drives the diaphragm. A small portion of the transmitter output is coupled to the reference cavity and detected by a crystal detector coupled to the cavity. If the transmitter is tuned to the mean frequency of the reference cavity, the crystal will detect resonance twice for every complete cycle of the modulation applied to the reference cavity.

Comparison of the phase of the crystal output with the phase of the modulating voltage shows that the fundamental of the modulating frequency in the crystal output will change in phase by 180° as the transmitter frequency shifts from one side of the mean frequency of the reference cavity to the other. The even harmonics do not change in phase under the same conditions. Discrimination is achieved by filtering out all but the first harmonic of the modulation frequency from the crystal output and by comparing the phase of this voltage to the phase of the voltage applied to the modulator of the reference cavity.

The amount of first harmonic present, and thus the amount of useful information available, is affected by the pulse width of the transmitter and the number of pulses present. The effective pulse width can be increased by allowing the pulse to charge through a low impedance a condenser that has to discharge through a high impedance. At low repetition rates there are few pulses, and thus few data from which to extract the first harmonic; this may cause an objectionable phase "flutter." Experience shows that, with sufficient pulse broadening, the phase flutter is not objectionable if the repetition rate is at least three times the modulating frequency. Since the pulse-repetition frequencies of interrogators usually exceed 200 cps, 60-cps modulation is satisfactory.

The modulated-cavity system gives trouble when used in a beacon with a receiver that has a square-wave-modulated local oscillator. Signals received over most of the band are modulated at the switching frequency, which is about 150 to 200 cps. Beats between the pulse-repetition frequency and the switching frequency may introduce low-frequency components which interfere with proper operation. Because of these difficulties, the two-crystal standing-wave discriminator has met with greater favor.

*Two-crystal Standing-wave Discriminator.*—The two-crystal[1] methods of discrimination operate on the principle that a cavity presents a capacitive susceptance above the resonant frequency, and an inductive susceptance below resonance. To make use of this fact, a portion of the r-f

---

[1] Another type of two-crystal system, not discussed further here, uses a magic T as a discriminator. A comprehensive treatment of the magic T as an r-f discriminator is given in *Technique of Microwave Measurements*, Vol. 11, Chap. 2, Radiation Laboratory Series.

energy is taken from the transmitter output through a directional coupler into a section of waveguide terminated by a sealed reference cavity of the desired frequency. When this is done, the standing waves set up in the waveguide by the reference cavity termination will change in phase and amplitude as the transmitting frequency shifts from one side of the resonant frequency of the cavity to the other.

Two pickup crystals are located in this waveguide. One is placed an odd number of eighth wavelengths from the equivalent cavity position, the other is spaced an odd number of quarter wavelengths from the first crystal. The two crystals will accordingly be at maxima of the standing-wave pattern opposite in phase when the frequency corresponds to the cavity resonance. At frequencies off resonance, the output of one crystal will exceed that of the other, the difference will change sign as the frequency passed through resonance, so that the discriminator characteristic of Fig. 13·9 will be obtained. The standing-wave patterns are shown in Fig. 13·10.

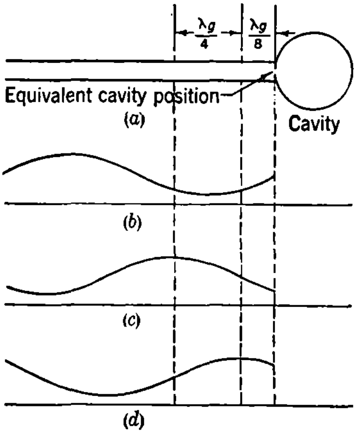


FIG. 13·10.—Standing-wave patterns in two-crystal standing-wave discriminators. (a) Position of crystals with respect to resonant cavity. (b) Standing-wave pattern at resonance. (c) Standing-wave pattern off resonance in one direction. (d) Standing-wave pattern off resonance in the opposite direction.

Initial balance of the crystal outputs when the reference cavity is at resonance is secured by gain controls in the amplifier stages so that zero d-c voltage is derived from balanced diode detectors. The balance is destroyed by a shift of the frequency to one side of resonance and the d-c output-control voltage increases positively in one channel and negatively in the other. A shift of frequency to the other side of resonance produces a d-c control voltage with reversed polarity. The discrimination characteristic is thus obtained by taking the difference of the two rectified output voltages.

The loaded $Q$ of the cavity controls the slope of the discriminator characteristics of Fig. 13·9. The useful range of control is the frequency range over which the characteristic is nearly straight. High-$Q$ cavities will give sharp slopes and better frequency control at the expense of a smaller useful range of control. The loaded $Q$ of the cavity may be adjusted to give the desired width of frequency range covered between discriminator points either side of center frequency. Most sensitive control of frequency will occur when the cavity is matched to the characteristic impedance of the line.

**13·12. AFC Amplifiers.**—The purpose of the AFC amplifier is to take error information from the discriminator and amplify it enough to actuate the motor driving the frequency-changing mechanism. The features of amplifiers of special importance for AFC of beacons will be discussed in this section.

Error information comes from the discriminator in the form of pulses. The amplifier must retain the amplitude information of the pulses; it is also desirable to stretch the pulses in order to increase the average current. Pulses are stretched by charging a condenser through a low resistance by means of a tube and letting it discharge through a high resistance before the next pulse arrives.

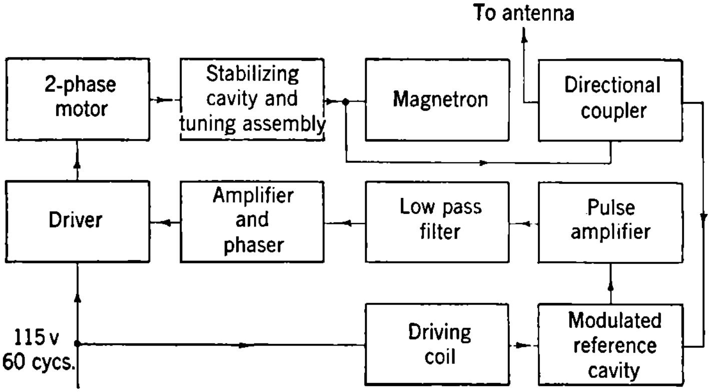


FIG. 13·11.—Block diagram of modulated-cavity method of AFC.

One problem that must be solved in servomechanism systems of this kind is that of "hunting," or overcorrection by the motor. When a correction has been completed and the motor switched off, the motor coasts through the position corresponding to the correct frequency and tunes the system off frequency in the opposite direction. An error voltage of opposite sign is immediately created and the motor reverses its direction. This cycle continues and causes the system to hunt.[1] Although the hunting can be eliminated by reducing the amplifier gain, the sensitivity is also reduced. Automatic gain control is a satisfactory solution; it has the advantage of not affecting weak signals seriously.

*Amplifier for the Modulated-cavity Discriminator.*—The error information from the modulated-cavity type of discriminator is the phase shift and change of amplitude in the first harmonic of the modulation of the

---

[1] The theory of servomechanisms of this type is treated in detail in *Theory of Servomechanisms,* Vol. 25, Chap. 4, Radiation Laboratory Series.

reference cavity. The amplifier uses this information to control the grids of a double triode which, in turn, supplies excitation for one phase of a two-phase a-c motor.

The amplifier circuit consists of a stage of pulse amplification followed by a low-pass filter, which passes only the first harmonic of the modulating frequency. Three stages of resistance-coupled amplification increase the level of this signal before it is applied to the grids of the driver which governs the direction of rotation of the motor. A gain control in the final amplifier stage, and AGC in earlier stages, are used to control hunting. A block diagram is shown in Fig. 13·11.

*Amplifier for Two-crystal Standing-wave Discriminator.*—A block diagram is shown in Fig. 13·12. The two-crystal discriminator has two separate voltage outputs. One output voltage increases while the other decreases when the transmitter shifts away from the frequency of the reference cavity. The amplifier therefore has two channels of typical audio-frequency design, combining resistance and transformer coupling between stages. Diode rectifiers convert the amplified pulsed information to d-c control voltages. The diodes are balanced for zero output at the beacon frequency by adjusting the gain. Sufficient amplification is available to permit the crystals to work at very low level, a total of 70 db attenuation being used between the r-f pickup and the crystal input. It is not found necessary to change crystals nor to make other adjustments in the system over long periods of continuous operation.

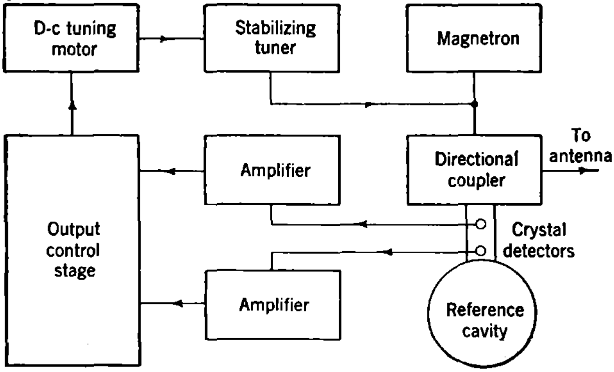


FIG. 13·12.—Block diagram of two-crystal AFC.

Automatic gain control is applied by taking a portion of the d-c voltage developed across the diode load resistor as a bias for the input to the first stage of the twin amplifier triode tube. The negative voltage developed in the diode rectifier of one channel is used as a bias in the input circuit of the first amplifier stage of the other channel to reduce the positive voltage at the output of the rectifier in this channel. Strong signals are thereby prevented from overdriving the motor; weak signals are affected only slightly. A potentiometer, as part of the diode load resistor, is used for final adjustments.

**13·13. AFC Servomechanisms.** *Modulated Cavity System.*—The servomechanism for the modulated cavity can best be described by analyzing the simplified circuit of the driver and its effect upon the frequency-correcting motor. The motor used in this system is of the two-phase "squirrel-cage" induction type. The direction of rotation can be reversed by retarding or advancing the phase of the current in either field winding by 180°. A method of accomplishing this is shown in Fig. 13·13. In this figure $F_1$ and $F_2$ are the two field windings of the motor. The two capacitors $C_1$ and $C_2$ set the phases of the fields approximately 90° apart. The combination of the inductance of $F_2$ and the capacitor $C_2$ resonates at approximately the supply frequency and thus reduces the frequency distortion due to the rectifying effects of the tubes, resulting in practically pure first-harmonic current through $F_2$. The phase of this current depends upon which tube conducts more. If the phase of the signal applied to the grids is 90° out of phase with the voltage applied to the plates, the resultant field averaged over one cycle is zero.

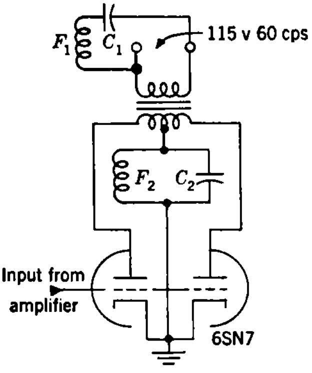


Fig. 13·13.—Driver circuit for two-phase servo motor.

*Two-crystal Standing-wave System.*—The most satisfactory servomechanism for the two-crystal standing-wave AFC system uses a small d-c motor, the field windings of which are excited by the plate currents of a twin-triode driver. The motor is designed with two field windings that can be used separately to govern its direction of rotation. By placing one field winding in the plate circuit of one triode unit and the other field winding in the plate circuit of the other triode unit, direct control of the motor in response to signals on the grids of the twin-triode tube is obtained. The armature is excited by a constant current.

In the two-crystal standing-wave AFC system, a d-c error voltage is developed at the grids of a twin-triode control tube in the final stage when the frequency is off resonance. There is no net error voltage when the beacon is on frequency, and therefore no signal voltage at the grids of the control tube. As soon as the frequency shifts, however, one or the other of the grids is driven positive and the corresponding plate current of that section causes the motor to rotate in such a direction that the system is retuned to the correct frequency. A 6SL7 tube is suitable for this application. Considerable degeneration is provided by a resistor in the common cathode return.

Limit switches or other preventive methods must be used to keep the motor from driving the tuning mechanism too far should the AFC lose control. The arrangement used in the servomechanism consists of a gear drive that moves the tuning plunger into or out of the cavity to correct for frequency changes. Reversing switches are actuated by the plunger at the limits of its travel. The reference cavity is designed to give a correction range of 12 Mc/sec on either side of the center frequency. Should the correction go beyond these limits, the drive continues until "pull-in" is again established on the return movement. This is accomplished by placing the plunger at the correct distance in the cavity during the initial adjustments on beacon frequency and making the limits of the movement such as to cover the tuning range desired. The reduction-gear box is designed to give a tuning rate of about 2 Mc/sec$^2$, which is satisfactory for all operational requirements.

*AFC with Short Interrogations.*—When a single scanning radar interrogates a beacon at long range, the beacon may not be interrogated for a long enough period to permit the AFC system to make a full correction. A device called "Yehudi" has been designed to overcome this difficulty. Yehudi begins its cycle of operation immediately after the first interrogation. The first interrogation closes a relay in the AFC unit. A set of contacts on this relay then causes a relay in Yehudi to close and to hold itself closed even after the AFC relay opens. The relay in Yehudi starts a 1-rpm motor which closes a pair of contacts for about 3 sec. Closing these contacts sets off a 400-cps blocking oscillator which trips the beacon at this rate for 3 sec, after which the oscillator stops. The motor continues to turn for another 57 sec, after which it automatically opens the relay, thus turning itself off. Yehudi is now set to repeat its cycle if the beacon is again interrogated. The 3-sec steady interrogation is sufficient to pull the beacon to correct frequency. Yehudi thus tunes

TABLE 13·1.—AFC PERFORMANCE DATA

| Frequency shift, Mc/sec | | | Time for shift and correction, sec | |
|---|---|---|---|---|
| PRF change | No AFC | AFC on | No AFC | AFC on |
| 200 → 400 | 9310.0 → 9309.0 | 9310.0 → 9309.8 → 9310.0 | 40 | 2 |
| 200 → 600 | 9310.0 → 9309.0 | 9310.0 → 9309.5 → 9310.0 | 30 | 1 |
| 200 → 800 | 9310.0 → 9308.3 | 9310.0 → 9309.3 → 9310.0 | 30 | 1 |
| 200 → 1000 | 9310.0 → 9308.0 | 9310.0 → 9309.3 → 9310.0 | 28 | 1 |

the beacon when it is interrogated and maintains silence when no one is using the beacon.

*Performance Data.*—Typical performance data on a 3-cm ground beacon, showing comparisons between the operation of the beacon without AFC and with a two-crystal standing-wave AFC system, are presented in Table 13·1. The magnetron was thermostatically temperature-controlled. It was tuned by a stabilizing tuner. The data are for a code of six pips, each of 0.5-μsec duration, the beacon being triggered by a signal from a 3-cm test set. The beacon was manually tuned to the correct frequency, 9310.0 Mc/sec, at a 200-cps repetition rate.

# CHAPTER 14

# TRIODE TRANSMITTERS

By P. A. de Paolo, K. R. More, and J. C. Reed

## PROPERTIES OF TRIODE OSCILLATORS

By P. A. de Paolo

**14·1. General Considerations.**—Triode oscillators have been used in beacon transmitters at frequencies up to 3300 Mc/sec. Triodes suitable for use at higher frequencies are not currently available, nor are suitable magnetrons available at the lower frequencies. Magnetrons are available for radar use at frequencies below 3000 Mc/sec, but they have not been used in beacons until now (1945).

A triode oscillator is somewhat more flexible than a magnetron in several respects. For example, a triode may be made to perform satisfactorily over a plate-voltage range of 5 to 1 or more (except when transit-time limitations intervene), without any circuit complications; this cannot be accomplished readily with a magnetron. A given triode may be made to oscillate over a very wide range of frequencies by the proper choice of the external circuit. On the other hand, a magnetron is designed for a given frequency range and the tunable magnetrons now available can be operated only over a narrow range ($\pm 5$ to 10 per cent) of frequencies around the design center.

Magnetrons, in general, are more expensive to build than low-power triodes,[1] when the two are compared on a power basis. Although this may not be true for high-power units, no comparison can be made because high-power triodes have not been built for use in the regions in which magnetrons are used.

In recent years, magnetrons have been improved rapidly because radar applications required very high pulse power. This concentration on magnetrons has resulted in a relative lack of attention to the development of triodes for use in the uhf and microwave regions.

At the present time few transmitter tubes having a power output above 5 kw are available for beacon use in the frequency range from 500 to 3000 Mc/sec. Development of such tubes is required in order to fill

---

[1] A more correct comparison is, perhaps, the cost of the magnetron with the cost of the triode and its associated oscillator circuit; however, in replacements, only the triode is replaced.

the need for beacons having power outputs above 5 kw in the uhf range above 500 Mc/sec. One of the most promising tubes available for this frequency range is the 2C39 ("oil can") tube. Even this tube, however, when used in a push-pull circuit, is limited to a maximum power output of about 10 kw. Furthermore, since it is difficult to utilize push-pull operation in cavity circuits, even this power is probably difficult to achieve above 1000 Mc/sec. Attention should be given to the development of small magnetrons for beacon use in this region.

It is well known that special precautions must be taken in the design of tubes and circuits for uhf and microwave use because lead length and transit time become important factors at these frequencies. The problem of heat dissipation also requires special attention because the tube elements are small.

**14·2. Basic Circuits.**—The triode transmitters used in beacons can be classified as follows:

1. Lumped-constant transmitters.
2. Parallel-line transmitters.
3. Coaxial-line transmitters.
4. Cavity transmitters.
5. Special (combinations of above).

Triode transmitters using lumped constants in the tank circuit are conventional and have been used as pulsed-beacon transmitters, either single-ended or in push-pull circuits. Principles that must be observed in designing this type of circuit for pulsed operation will be discussed in Sec. 14·8.

*Parallel-line Oscillators.*—Parallel-line oscillators in which the lumped inductors and capacitors used in the resonant circuits are replaced by parallel transmission lines have been used in beacon systems. The parallel lines may be looked upon as having small distributed elements of inductance and capacitance which function in combination like the normal "tank" circuit.

The lines are referred to as being one-quarter, one-half, three-quarters, etc., of a wavelength long because their lengths approach these dimensions when the lead length and tube interelectrode capacitances are small. In some cases, however, the tube capacitances are large, with the result that the actual length of quarter-wave lines may be only a fraction of a quarter wave.

Depending on the tube used and other circuit conditions, the oscillator tube may be placed either at the end of quarter-wave lines or at the center of half-wave lines. The choice depends largely on the frequency range of operation of a given tube. For example, suppose it is desired to operate a certain tube as a parallel-line oscillator near its upper frequency limit. The use of quarter-wave lines would require that the lines be very short; however, if the tube were placed at the center of half-wave lines, the effective line length would be increased. Thus it may be said that, if the frequency is low, the length of the line will be quite long and quarter-wave lines should be used. However, as the frequency increases and the quarter-wave lines become inconveniently short, it is advisable to use either half-wave lines or coaxial lines, assuming that a suitable tuning method is used.

Parallel-line oscillators are tuned by changing the length of the lines, by capacitance or, rarely, by inductance tuning. The former method is the most desirable. The length of the lines may be changed readily in push-pull circuits; this change is more difficult to accomplish in single-ended circuits. Capacitance loading at the high-impedance end of the lines is readily accomplished in either case. The tuning capacitor should have very low losses and excellent mechanical stability. Inductance tuning is accomplished by introducing a movable loop that decreases the inductance by acting as a short-circuited secondary.

*Coaxial-line Transmitters.*—Some transmitters use coaxial lines rather than parallel lines for the resonant elements. This has some advantages when tubes like the lighthouse tube (2C40, 2C43) which lend themselves readily to coaxial-line construction are used. Another advantage of coaxial-line circuits is that contact fingers rather than flexible leads are used to connect the tube in the circuit. Such contact fingers, in general, have larger area and lower inductance; the upper limit of the operating frequency will be increased, therefore, with the same interelectrode capacitance. Coaxial-line circuits are, in general, self-shielding and thus relatively free from radiation losses. This is important when space is limited, since shielding is usually necessary, except at the lowest frequencies. For a given tuning range, power output, and circuit efficiency, coaxial-line oscillators will occupy less space than parallel-line transmitters.

Tuning is accomplished as with parallel lines by changing the length of the resonant line or by capacitance loading at the high-voltage end of the line. Inductance tuning is not convenient. Tuning by changing the line length of quarter-wave coaxial oscillators is not difficult to accomplish; it is rather difficult, however, with half-wave lines.

*Cavity Oscillators.*—Oscillators that use cavities as resonant elements are referred to as "cavity oscillators." A quarter-wave coaxial-line oscillator used at ultrahigh frequencies is, in a sense, a cavity oscillator. Cavity oscillators take many forms but, in general, can be divided into two types, the reentrant and nonreentrant cavity.[1]

---

[1] For a discussion of reentrant and nonreentrant cavities the reader is referred to *Klystrons and Microwave Triodes*, Vol. 7, Chaps. 7 and 8, of this series.

The cavity resonator has high $Q$, high shunt impedance, and high efficiency. Cavities are often used at ultrahigh and microwave frequencies for beacon transmitters. They may be used at lower frequencies, but the cavities then become so large that designers often prefer parallel-line or coaxial-line oscillators. The size of the cavity at low frequencies can be reduced by "folding"; the r-f energy is then caused to travel over large surfaces by dividing the cavity into compartments. This has the same effect as increasing the size of the cavity.

At frequencies above about 500 Mc/sec, the designer must use coaxial lines or cavities for triode transmitters, since the elements of other types of circuit become so small at higher frequencies that they are impractical. More important, at ultrahigh frequencies and microwave frequencies the circuit elements themselves may well become efficient radiators; the circuit may radiate all of the energy it can supply without any connection to an antenna circuit if enclosed electrical circuits are not used. Conversely, at these high frequencies, the oscillator is sensitive to other components within its radiation field. Shielding is thus an important consideration; cavities are self-shielding. If the designer uses holes and slots in the cavity for ventilation, adjustment, etc., care must be taken to locate and orient them suitably, for they can become effective radiators at ultrahigh and microwave frequencies.

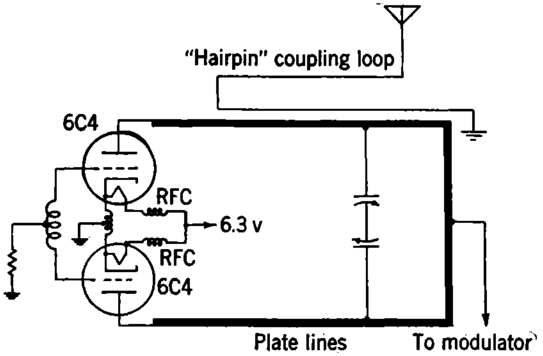


FIG. 14·1.—Miniature beacon transmitter at 200 Mc/sec using parallel lines in the plate circuit and lumped constants in the grid circuits. NOTE: inductance tuning by rotating a closed loop in the plate circuit.

Combinations of parallel lines and lumped-constant circuits have been used as resonant elements in beacon transmitters. For example, in Fig. 14·1 is shown a push-pull transmitter using parallel lines in the plate circuit and lumped constants in the grid circuit. The cathode and filament circuits are isolated by means of r-f chokes. This combination was used for saving space, and in general this is probably the only reason for the use of such a combination.

Other combinations of lines, cavities, and lumped constants can be used in the same oscillator circuit.

**14·3. Frequency Range and Limitations.**—The basic transmitters discussed in the previous section have fairly definite frequency limitations and are used only over certain frequency ranges for reasons of size, efficiency, and simplicity.

There are certain limitations that apply in greater or lesser degree to all the types of oscillators discussed. One of these is the plate voltage, whether it is d-c or pulse voltage. Tubes designed for uhf and microwave operation have small elements closely spaced to reduce transit-time effects. These small spacings limit the plate voltage that may be used and consequently limit the output power. This limitation becomes more and more serious as the operating frequency is increased.

A second limitation that confronts the designer, and becomes increasingly serious at higher frequencies, is tube dissipation. As was pointed out above, the elements in high-frequency triodes are small and therefore limited as to the power that they can dissipate.

Another limitation that may cause trouble for plate-pulsed oscillators is a low $L/C$ ratio in the plate circuit. The capacitance of the tuned circuit is in shunt with the pulse circuit. When this capacitance is large, input pulses having short rise-and-fall times will suffer deterioration.

Design characteristics that will correct some of the limitations above conflict with steps taken to correct others. For example, when the tube designer cuts down the element spacing in order to decrease transit-time limitations, the voltage that may be applied to the tube will be decreased.

The circuit designer should select a transmitting tube that will deliver the desired power near its maximum voltage rating, with all other factors considered. In so doing, several desirable conditions are achieved. First, the tube selected will have the minimum size and heater drain for the application. Second, since the tube is operated with nearly maximum plate voltage, transit-time effects will be minimized. Often these conditions cannot be met, particularly at the higher frequencies, because the design may fall in a region in which a tube even approximately suitable for the application is unavailable.

*Lumped-constant Transmitters.*—Transmitters using lumped constants have been used in beacon systems up to frequencies in the region of 200 Mc/sec. The practical limit is currently in the region of 300 Mc/sec, and depends on the stability desired and on the tuning range the transmitters are to cover. The efficiency of this type of transmitter may be very poor at higher frequencies; extra precautions must, therefore, be taken in the selection of low-loss insulation for the support of the resonant elements. Changes in the physical dimensions of the inductance and capacitance elements due to temperature changes may shift the operating frequency over a wide range.

*Parallel-line Transmitters.*—Parallel-line transmitters have been used in beacons over the frequency range from 150 to 500 Mc/sec. These are not the frequency limits for this type of construction. The upper limit is determined by the triode used, by the tuning range desired, and by radiation losses. In general, it is desirable to keep most of the frequency-determining elements external to the tube. The lowest frequency at which lines are used, in general, depends on the space that can be assigned to the beacon transmitter.

*Coaxial-line Transmitters.*—Coaxial lines may be used at somewhat higher frequencies than parallel lines for three reasons: (1) tubes well adapted to coaxial-line construction generally have small lead lengths; (2) coaxial-line construction, in contrast to parallel-line construction, involves no flexible leads; (3) radiation losses are largely eliminated. Coaxial lines as resonant elements have been used at frequencies up to 1000 Mc/sec. The upper limit again depends on the tube used; for tubes currently available, the upper limit is probably in the region of 1200 Mc/sec. The lower limit depends on the space that the designer can devote to the transmitter. While coaxial lines have been used close to 1000 Mc/sec, it is probably safe to say that effective use in this region is the exception and not the rule, and that cavities are more stable, more efficient, and probably more satisfactory at these frequencies.

*Cavity Transmitters.*—The cavity-type oscillator has been used as a beacon transmitter at frequencies up to 3300 Mc/sec. Again the limiting factor is not the cavity but the triodes available for use at these frequencies. At the present time, the lighthouse tube (2C40 and 2C49) is the only type of triode that has been used in the field as a beacon transmitter at these high frequencies.[1] It must be remembered that the triode transmitters used at 3000 Mc/sec are usually low-power units—from 50 to 2000 watts pulse power—and that the magnetron is usually used for higher-power outputs.

**14·4. Delay and Impedance.** *Pulse-delay Characteristics.*—When triode oscillators are pulsed with rectangular pulses, the transmitted r-f pulse is delayed with respect to the modulator pulse. The magnitude of this delay varies with the type of oscillator used and with the rise time of the modulating pulse, as well as with the type of pulsing used. In general, the delay between modulator and r-f pulse is undesirable, and every effort is made to keep this delay to a minimum.

The factors affecting this delay are the rise time of the modulating

---

[1] However, a new tube that appears to be satisfactory in the region of 3000 Mc/sec has been developed. This is the Sylvania type SB-846B. Whether or not the SB-846B can be used at frequencies above 3000 Mc/sec has yet to be determined.

pulse, energy storage in the oscillator, the type of pulsing used (grid-, plate-, or cathode-pulsing), and the magnitude of the pulse voltage.

In the design of triode transmitters it is difficult, and in most cases impossible, to predict just what the delay will be. The designer can observe certain rules, however, which will simplify his problem and avoid unreasonable pulse delays. Long time constants in the grid circuit of the transmitter should be avoided. The use of degenerative circuits in the cathode of the transmitter should be avoided; they result in delayed pulses and lowered efficiency. If the transmitter resonant circuit stores much energy, one can expect to get pulse delays. This is especially true if the modulating pulse has a long rise time. In general, however, even though the resonant circuits by themselves may have a high $Q$, when the oscillator tubes are added and an antenna load is coupled tightly into the transmitter, the resultant loaded $Q$ will be considerably less than the unloaded circuit $Q$.

Other things being equal, the designer should try to use as high voltage as possible on the transmitting tube, especially if the tubes used are operating near the limit set by the transit time.

*Impedance-matching.*—In order to obtain maximum power from the r-f oscillator it is important to match the output impedance and, usually, the input impedance as well. The input impedance of the transmitter is often, though not always, matched to the output impedance of the modulator. More important for maximum power output[1] the output impedance of the oscillator should match the load impedance. Input impedance matching is especially necessary when a gas-filled tube modulator is used to plate-pulse the transmitter. When the transmitter impedance is different from the modulator impedance, the pulse will suffer reflection, and the main pulse may be followed by a second undesirable pulse. Should the transmitter impedance be lower than that of the modulator, the voltage at the r-f oscillator will be low and the power put out may decrease.

It is difficult to estimate the impedance of an oscillator intended for pulse operation unless the designer has had previous experience with the circuit or has data available as to the impedance of the tube as an oscillator. The pulse impedance increases as the plate voltage on the tube is decreased. This is not a linear relationship; the impedance varies as some power of the plate voltage, in the region that the cathode emission is not limited. The factors that influence oscillator impedance for a given tube are plate voltage and grid-circuit impedance. In general, most of the triode transmitters used in beacon applications have an impedance somewhere between 500 and 4000 ohms.

---

[1] It may sometimes be desirable to sacrifice power by mismatch to achieve greater frequency stability, etc.

Once a model of the oscillator is constructed, the input impedance may be determined by the substitution method. The proper operating voltage is applied to the oscillator from a source of variable pulse voltage. This voltage is observed with a synchroscope and a suitable voltage divider. The oscillator is then replaced with a variable noninductive resistor, and its value adjusted until the same pulse voltage is observed. This value of resistance equals the oscillator impedance. When more accurate results are desired, the pulse current may be measured by observing the voltage drop across a resistor in series with the oscillator supply voltage. The pulse supply voltage is also observed. From these two quantities the oscillator impedance may be computed. The impedance will vary as various oscillator tubes are substituted in the circuit. A spread between maximum and minimum impedance of about ±20 per cent for most triodes of a given type is to be expected.

## TYPES OF MODULATION FOR TRIODE OSCILLATORS

The triode transmitter may be modulated by means of plate pulsing, grid pulsing, or cathode pulsing; combinations of these three methods are also possible.

The type of modulation selected depends almost entirely upon the requirements placed on the beacon. For example, when accurate range information is needed, plate or cathode pulsing should be used to keep the delay time to a minimum constant value. On the other hand, if the beacon must be very light and a gas-filled tube modulator cannot be used, grid pulsing may have to be employed at the expense of extreme range accuracy. Grid pulsing should not be used, however, if the beacon is to be coded and the time between pulses is comparable to the pulse width.

In general, one may say that if the delay time is to be kept to a minimum and the amount of driving power is no limitation, plate pulsing should be used. If the driving power must be kept to a minimum with low plate voltage, and delay time is not important, then grid pulsing will be the favored method. Cathode pulsing may be used when the available driving power is limited, but the delay time is to be kept within the limits obtainable with plate pulsing. Detailed discussion of the various methods follows.

**14·5. Plate Pulsing.**—Plate pulsing, which is most widely used in beacon systems, has several advantages over other pulsing methods. One is that the delay time usually can be held to a minimum. Once the circuit has been designed to operate under a given set of conditions, the delay time will not vary with normal changes in plate voltage, heater voltage, and temperature, nor with the use of different tubes in the circuit. This, of course, is very desirable when precise range information is important because, if the delay is known and fixed, it can be readily accounted for in the associated circuits.

In the case of plate pulsing, no steady voltage is used on the oscillator. The transmitter is caused to oscillate by the application to the plate of a rectangular pulse of sufficient amplitude to give the desired power output and of such a width and shape that it will give the desired r-f pulse.

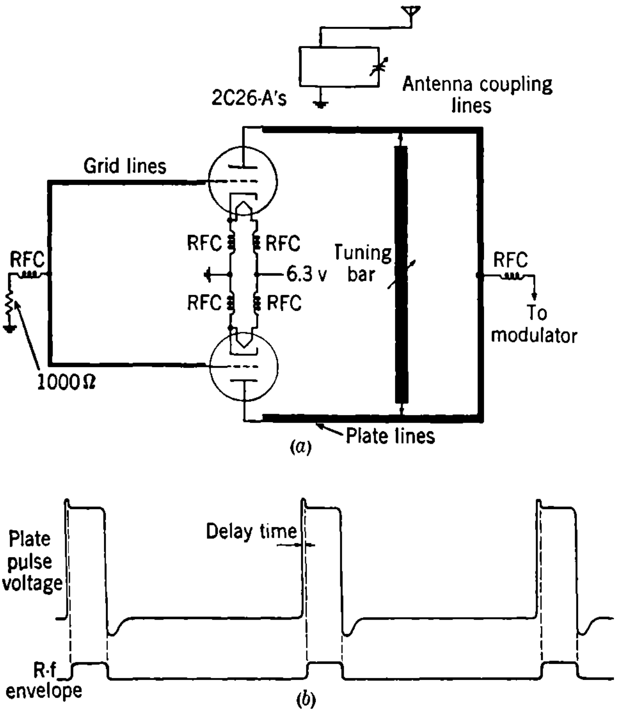


FIG. 14·2.—Circuit diagram and typical waveforms for a push-pull triode transmitter at 200 Mc/sec, suitable for plate pulsing.

Another advantage of plate pulsing over other methods is that a lower d-c supply voltage may be required for the same power output since the pulse transformer may have a stepup ratio. Designers have also found that tubes can operate at higher plate voltages when they are plate-pulsed than when they are grid-pulsed. For example, the 2C26A is rated at 3500 volts maximum on the plate of the tube for plate-pulsed conditions, but it is rated at only 2500 volts maximum for grid-pulsed conditions.

Plate pulsing is very satisfactory when the beacon is to be coded because the transmitter circuit recovers very rapidly after the pulse is over.

In Fig. 14·2 is shown a plate-pulsed triode transmitter. It will be noted that no d-c voltages are used.

**14·6. Grid Pulsing.**—Grid pulsing differs from plate pulsing in that a constant d-c potential is applied to the plate of the oscillator. The grid bias is generally fixed and applied to the grid circuit, although cathode bias has been used. The bias is fixed at a value such that the transmitter tube will not oscillate during the quiescent period. The modulating pulse drives the grid of the oscillator tube or tubes into the region in which the tube oscillates.

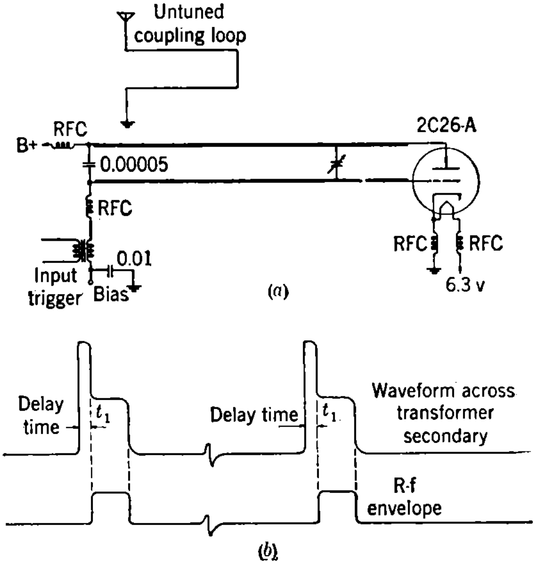


Fig. 14·3.—Circuit diagram (*a*) and waveforms (*b*) for grid-pulsed oscillator.

The bias level is determined by the d-c voltage applied to the plate of the tube, and the bias is made high enough so that any tube placed in the circuit will not oscillate unless the proper modulation pulse is applied to the grid. For any given plate voltage, the power developed by a grid-pulsed oscillator will increase as the grid drive is increased. The limits are reached when the grid drive reaches a point at which sparking occurs between grid and cathode, when the drive is increased to the point at which the maximum grid dissipation of the tubes is reached, or when the peak cathode emission becomes the limiting factor.

Grid pulsing is often used when a single-pulse reply code is to be transmitted and the d-c voltage necessary to give the required power is low (under 1200 volts or thereabouts). By using grid pulsing, the modulator is made very simple since the power that the modulator has to deliver to the grid of the oscillator is small compared with the power that the modulator would have to deliver to the plate were the transmitter plate-pulsed.

Grid pulsing, however, has disadvantages. One is the variable delay between the applied modulator pulse and the r-f pulse. This delay time varies with grid bias, grid drive, plate voltage, tube characteristics, and with antenna loading. It is designated as $t_1$ in Fig. 14·3, which shows the circuit diagram and waveforms for a grid-pulsed oscillator.

By observing a few rules, designers can keep the delay variation to a minimum. The modulating pulse should have a fast rise time and the time constant of the grid circuit should be as short as possible so that the leading edge of the modulating pulse will not deteriorate. Grid bias is more desirable than cathode bias, not only to reduce pulse delay, but also to eliminate cathode degeneration. It is desirable to regulate the grid-bias supply voltage.

In general, grid pulsing with techniques now in use appears less desirable than plate or cathode pulsing. When either of these is used, the grid can be tied to r-f ground and much trouble avoided. Furthermore, experience tends to indicate that plate-pulsed tubes have longer life, although no thorough study has been made of tube life under various conditions. The currently accepted explanation for the discrepancy in tube life is that the plate pulses used are short; accordingly, ions formed by ionization of residual gas in the tube do not move far in the short period of the pulse. In consequence, arc-over and sparking within the tube do not often get started, and do not cause permanent damage even if they do start. With grid pulsing, on the other hand, any transient that causes the tube to arc over will result in dissipation of damaging amounts of energy in the tube because of the steady d-c plate voltage.

**14·7. Cathode Pulsing.**—In Fig. 14·4 are shown a circuit and waveforms for a transmitter using cathode pulsing with a fixed d-c voltage on the plate of the oscillator. The capacitor used in the plate supply is large enough to ensure that the power-supply impedance will be very low during the pulse interval. In this circuit, the grid is tied directly to the r-f ground and the bias is applied in the cathode circuit. Enough bias is used to insure that the tubes are practically biased to cutoff during the quiescent period. The negative modulating pulse is applied to the cathode circuit by means of a pulse transformer. During the pulse interval, the cathode is driven negative with respect to the bias level and the transmitter is caused to oscillate actively for the pulse interval. As with grid pulsing, power output from the oscillator will increase with increase in the drive voltage. When cathode pulsing is used, the delay between the modulating pulse and the r-f pulse is comparable to the delays caused by plate pulsing. This delay is usually small, and fairly constant over wide ranges of plate voltage and bias voltage. It does not vary much from tube to tube.

Several precautions must be taken to achieve good design. Since the winding capacitance of the filament transformer is in shunt with the secondary of the pulse transformer, it must be kept low if pulses with a fast rise are supplied by the modulator. When cathode-type tubes are used, this is not so important, for the cathode-heater capacitance will usually be small. Unless the heater winding is allowed to "float," however, the designer must make sure that the cathode bias required to cut off the oscillator does not exceed the maximum rated heater-cathode potential.

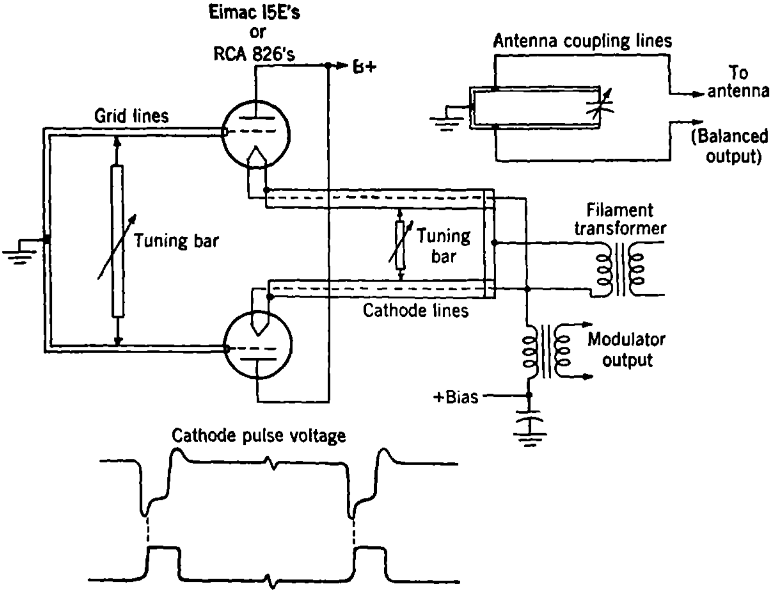


FIG. 14·4.—Circuit diagram and typical waveforms for a 170-Mc/sec cathode-pulsed oscillator.

The designer must also make sure that the pulse impedance in the cathode circuit of the oscillator is kept to a minimum. This means that the pulse transformer must have a low-impedance secondary winding and that a large bypass capacitor must be used in the bias supply. As the cathode is driven more and more negative with respect to the bias level, the impedance of the cathode circuit decreases. This means that for high power output from the transmitter, corresponding to a high driving voltage from the modulator, the modulator is required to supply a considerable amount of power to the cathode circuit of the transmitter

*Combination Pulsing.*—A combination of plate and cathode pulsing has been used in several portable beacons in order to realize the desired power at a low d-c supply voltage. In general, combination pulsing is not desirable because it adds complications to the system and, except in rare instances as mentioned above, has no particular advantages.

*Comparison of Pulsing Methods.*—In general, when space and power are available, plate pulsing is preferable to the other methods, since it is the easiest to handle. Grid pulsing may be considered when the modulator must be simple and the plate voltage necessary for the desired power output is low. When the modulator can supply some power, but not enough to accomplish plate pulsing, and a constant delay time is desired, cathode pulsing should be considered.

## VHF AND UHF TRIODE OSCILLATORS (200 TO 1000 MC/SEC)

In this section a few oscillators that have been used in beacons will be discussed in a general way to illustrate a few of the problems involved in the design of oscillators for beacon service.

The oscillators to be described use conventional lumped-constant elements, parallel lines, coaxial lines, and cavities as resonant elements. They cover a frequency range from 200 to 1000 Mc/sec and have power outputs from 10 to 3000 watts of pulse power.

**14·8. Lumped-constant Oscillators.**—As we have seen, the practical upper limit of frequency for this type of oscillator is in the region of 300 Mc/sec for the tubes normally used.

The efficiency for pulsed operation of this type of circuit is 20 to 30 per cent in the region of 200 Mc/sec for the tubes currently used in beacon service. It is determined largely by the quality and design of the components used in the oscillating circuits. It is important, therefore, that the designer choose these components carefully and integrate them into the circuit so as to make the tube efficiency be the limiting factor.

**14·9. Parallel-line Oscillators.**—Parallel-line oscillators have been used for beacons in the frequency range from 150 to 500 Mc/sec and at power levels from 10 to 3000 watts. This type of transmitter is stable and can be tuned mechanically over a fairly wide frequency range without too much difficulty. For use at the lower frequencies this type of oscillator may be quite large, since the length of the resonant lines may approach one-quarter wave of the resonant frequency. Parallel-line oscillators are inferior to coaxial-line oscillators, however, because they require shielding.

Figure 14·5 is a photograph of a push-pull transmitter using tuned lines in the plate circuit, and untuned grid and cathode circuits. This transmitter was designed for operation in the region of 200 Mc/sec and may be tuned over a 5 per cent band. A wider band of frequencies

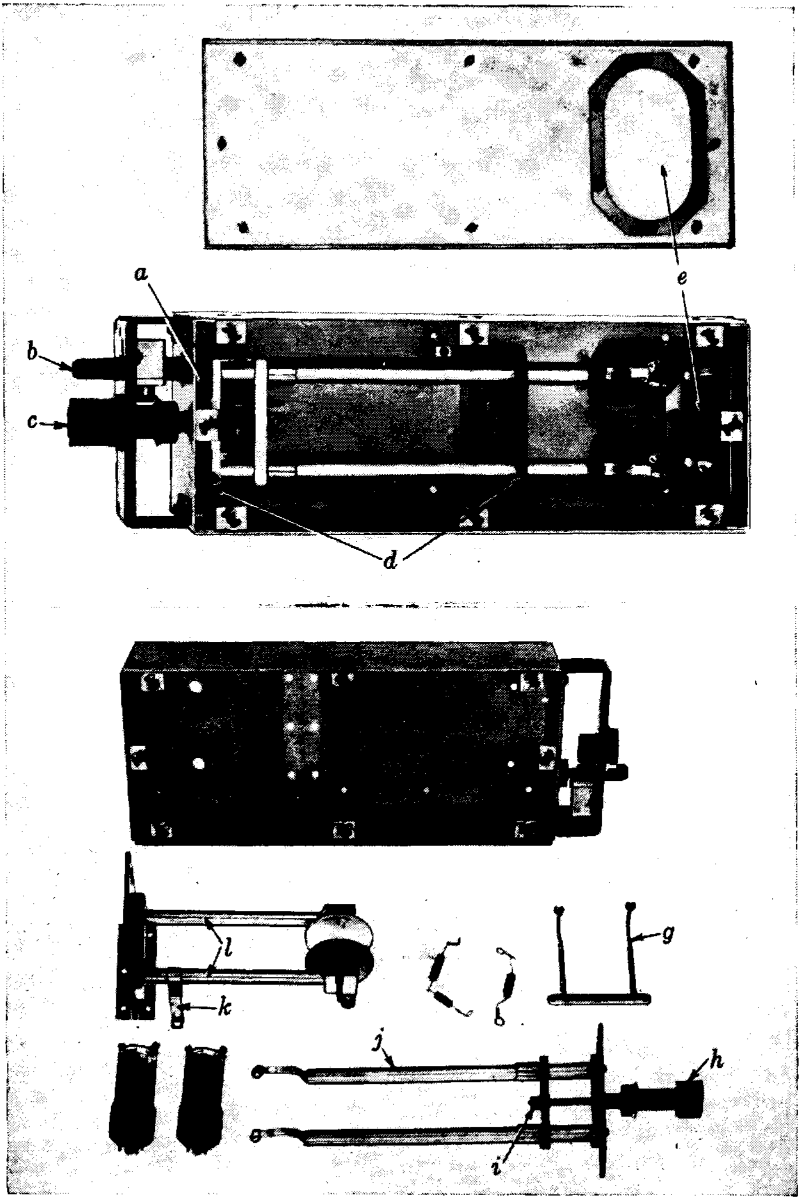


FIG. 14·5.—Photograph of a complete 200-Mc/sec beacon transmitter and its component parts. The circuit of this transmitter is shown in Fig. 14·2. (*a*) Frequency indicator (Veeder-Root counter). (*b*) Antenna tuning control. (*c*) Frequency control. (*d*) Insulating supports. (*e*) Ventilation grille. (*f*) Output connection. (*g*) Grid-circuit assembly. (*h*) Tuning knob and tuning mechanism (insulated). (*i*) Movable shorting bar. (*j*) Plate lines. (*k*) Ground strap. (*l*) Antenna coupling lines.

could be covered by lengthening the lead screw actuating the short-circuiting bar. This oscillator was designed for plate pulsing. It supplies 2.5 to 3 kw of pulse power when a plate voltage of 3000 to 3500 volts is used. The power output is essentially constant over the band when the transmitter is feeding a matched load. The plate-power efficiency is about 20 per cent and could be increased by increasing the antenna coupling. This, however, would make the system more sensitive to changes in line voltage and standing-wave ratio of the load. The frequency stability of this oscillator is ±0.13 Mc/sec for plate-voltage variation of ±10 per cent and temperature variations from −40°C to +40°C.

Figure 14·2, Sec. 14·5, shows the circuit diagram for the transmitter shown in Fig. 14·5. It will be noted that this oscillator utilizes quarter-wave lines in the plate circuit, and untuned lines in the grid circuit. Radio-frequency chokes are used to isolate the cathode and heater from the common r-f ground. The length of grid lines was chosen to give the desired grid excitation, and the size of the lines is such that the resonant frequency of the grid circuit is far removed from the frequency of operation of the oscillator. This is important for, if the resonant frequency of the grid circuit falls within the operating range of the oscillator, frequency instability may be encountered, with the transmitter tending to oscillate in an unloaded mode.

When space is available, it is desirable to use lines to tune the grid, plate, and cathode. Space limitations prevented this for the design under discussion, so it was elected to tune the plate circuit and isolate the cathode and heater circuits by means of low-loss r-f chokes.

The r-f energy is coupled out of the circuit by means of quarter-wave capacity-tuned lines. The oscillator output circuit is matched to the transmission system by tapping the lines so that the oscillator output impedance is equal to the impedance of the transmission system.

**14·10. Coaxial-line Oscillators.**—The coaxial line can be either one-quarter wave long or one-half wave long, depending on space requirements, the tube used, and the complexity of the tuning mechanism. A quarter-wave coaxial-line oscillator can be designed to have a very simple tuning mechanism, but the isolating capacitor between the center conductor and the tube presents a difficult mechanical problem. On the other hand, in the case of a half-wave oscillator, no blocking capacitor is required, but the tuning mechanism may become more complicated.

*The 500-Mc/sec Oscillator.*—In Figs. 14·6 and 7·14 is shown a half-wave coaxial-line oscillator using an Eimac 15E tube. It is plate-pulsed and intended for operation in the region of 500 Mc/sec. The pulse voltage is fed to the center conductor of the line at a point having approximately zero r-f potential and no blocking capacitor is necessary. An r-f choke is used at the point where the pulse voltage is applied so that no r-f energy is fed back into the pulse circuits. This choke is necessary because the oscillator tunes over a band of frequencies, and tuning is accomplished by increasing or decreasing the length of the center conductor. This means that the pulse tap can be at zero r-f potential for one frequency only and will be unbalanced for any other frequency. The filament circuit is bypassed by small built-in mica capacitors. Ordinary capacitors cannot be used at these frequencies because of lead inductance.

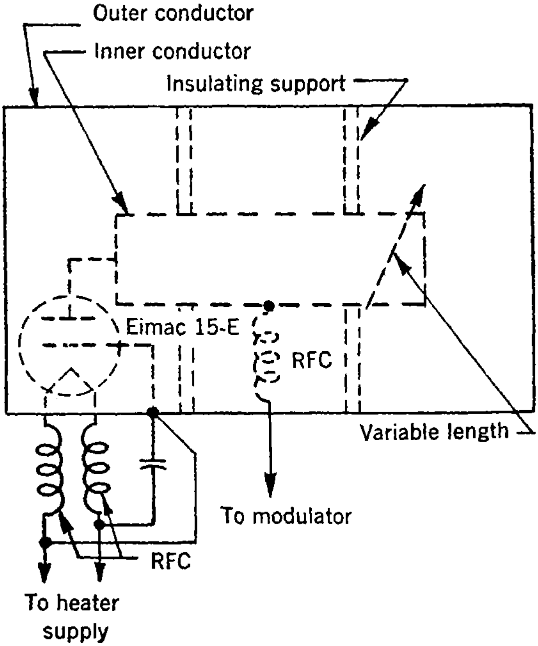


FIG. 14·6.—Diagram of a half-wave coaxial line oscillator at 500 Mc/sec.

This oscillator can be pulsed at voltages from 6 to 10 kv and will deliver up to 5 kw of pulse power. Its efficiency at the higher voltages is about 30 per cent. The r-f energy may be coupled out either by a capacity probe or by a small coupling loop. The method used will depend on the load and the power variation that can be tolerated over the frequency band.

*The 700-Mc/sec Oscillator.*—Figures 14·8 and 14·9 show a photograph and circuit diagram of a miniature half-wave coaxial-line oscillator designed for operation in the region of 700 Mc/sec. It is unusual in that it is used as both transmitter and receiver in a small beacon. During the reception period it is a superregenerative receiver; during the transmission period it is a plate-pulsed half-wave coaxial-line oscillator. The tube used is a 6F4 which has double leads for the grid and plate elements to reduce lead inductance.

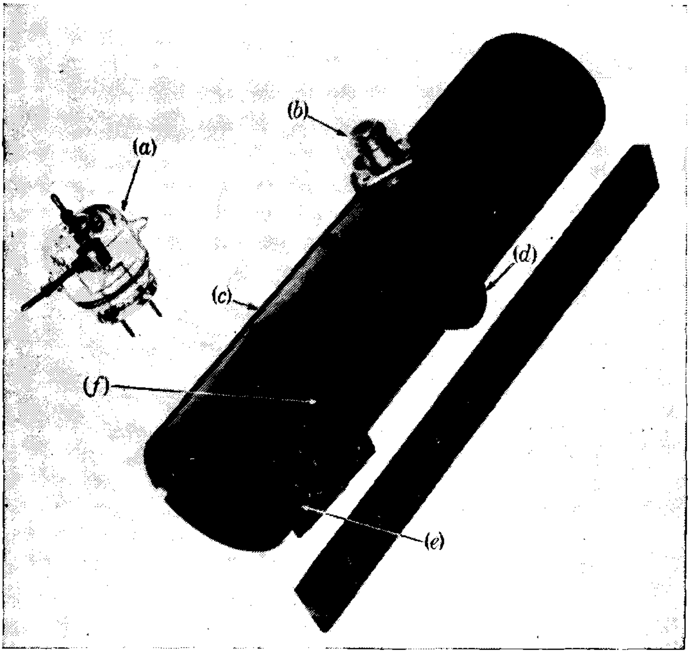


Fig. 14·7.—Photograph of the half-wave coaxial-line 500-Mc/sec oscillator shown schematically in Fig. 14·6. (*a*) Tube 15E. (*b*) R-f output connector. (*c*) Outer conductor. (*d*) High-voltage bushing for modulation voltage. (*e*) Filament connection. (*f*) Built-in mica capacitors.

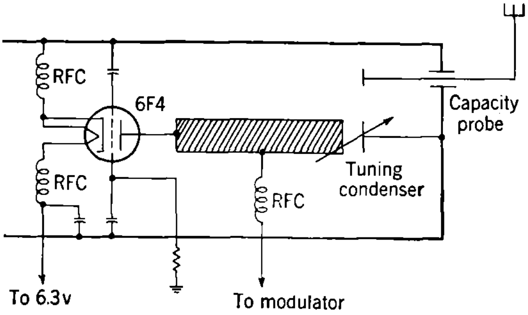


Fig. 14·8.—Miniature half-wave coaxial oscillator for operation at 700 Mc/sec.

Tuning is accomplished by changing the capacitance at the open end of the coaxial line. The photograph shows the effect of lead inductance and interelectrode capacitance of the tube on the physical length of the inner conductor of the line. The tuning mechanism on this circuit provides for tuning over about a 10 per cent band.

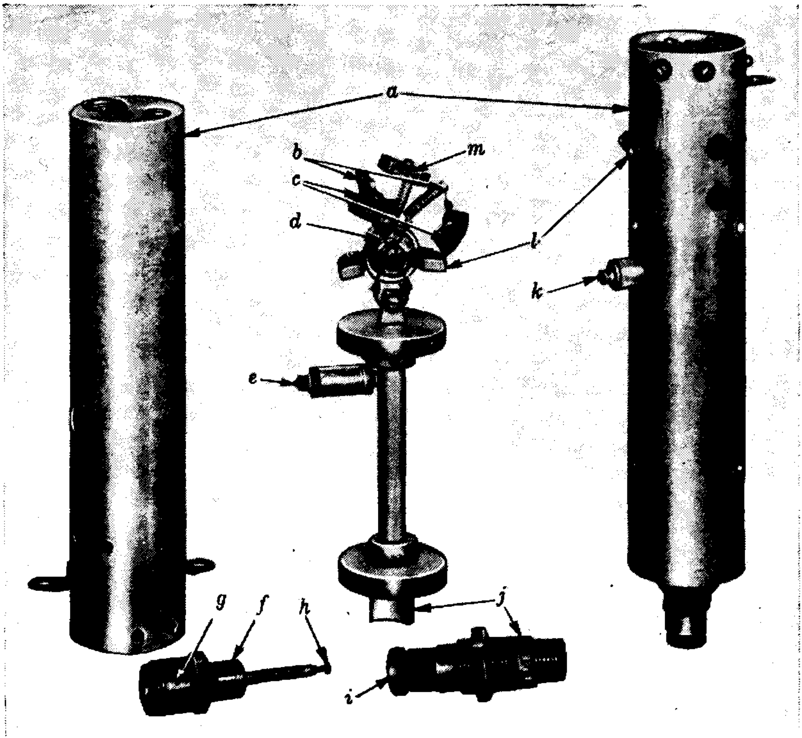


FIG. 14·9.—Photograph of a miniature half-wave coaxial-line 700-Mc/sec oscillator. The circuit diagram is given in Fig. 14·8. (*a*) Outer conductor. (*b*) Antenna connector. (*c*) Heater connector. (*d*) 6F4 tube. (*e*) Plate choke. (*f*) Capacity probe. (*g*) Antenna connector. (*h*) Probe-adjusting screw. (*i*) Micrometer tuning control. (*j*) Tuning capacitor. (*k*) Plate lead connection. (*l*) Grid connection. (*m*) Cathode connection.

Another unusual feature of this circuit is that connection clips and isolating chokes are soldered directly to the tube lead extensions; they are supplied as a part of the tube and are replaced with it.

As a transmitter, this circuit delivers a 0.75-$\mu$sec 15-watt pulse. The plate pulse voltage applied to the circuit as a transmitter is about 600 volts. Power is taken from the oscillator by means of a capacity probe coupled to the open end of the line (see Fig. 14·9). Micrometer adjustment of the coupling is obtained by means of a screw at the end of the probe. As a receiver, this circuit has a sensitivity of $2 \times 10^{-9}$ watt for triggering. Even though different voltages are applied to the plate of the tube during transmission and reception, the frequencies of transmission and reception differ by less than 2 Mc/sec. They can be made very nearly the same by careful adjustments, especially of the cathode impedance.

**14·11. Cavity Oscillators at 700 and 1000 Mc/sec.**—It was pointed out in Sec. 14·3 that, while circuits like coaxial lines may be used at ultrahigh frequencies, the cavity is especially suited for these frequencies.

Two examples of cavity oscillators used at ultrahigh frequencies are given below. Like the miniature half-wave coaxial oscillator mentioned above, both of these cavity oscillators are used as superregenerative receivers during the reception period and as plate-tuned cavity oscillators during the transmission period. Both cavity oscillators are provided with a coarse frequency adjustment so that they may be tuned to the center of the proper band. In addition, a motor-driven variable capacitor is provided. It sweeps the beacon frequency through a band so that the beacon may respond to interrogations on different frequencies in the band.

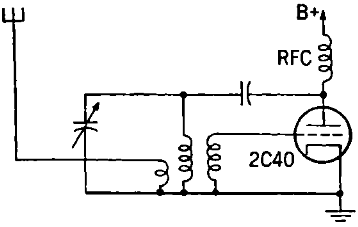


FIG. 14·10.—Equivalent circuit for the 700-Mc/sec cavity oscillator shown in Figs. 14·11 and 14·12.

*The 700-Mc/sec Oscillator.*—The cavity oscillator shown in Figs. 14·10, 14·11, and 14·12 is designed for operation in the region of 700 Mc/sec. It uses the 2C40 lighthouse tube and is pulsed in the plate and cathode circuits. The center conductor of the cavity is at the same r-f potential as the plate of the tube. A blocking capacitor is provided so that the d-c supply will not be short-circuited. Grid excitation is obtained by means of two inductive loops tapped in at the cathode end of the cavity. The cathode is connected directly to the cavity by a built-in bypass capacitor in the 2C40. The r-f energy is coupled out by means of an inductive, untuned coupling loop placed in the cavity at the plate end.

Coarse frequency settings are made by sliding the cathode end of the cavity in or out until it is approximately on frequency. The final adjustment is made by moving the rotor of the tuning capacitor with respect to the stator, which is fastened to the plate rod. The frequency is then swept by driving the rotor of the capacitor with a motor.

The coarse tuning covers about a 25 per cent band; the fine tuning covers a 10 per cent band. The frequency of the oscillator is swept over a 6 per cent band.

The pulse power output of this oscillator is about 50 watts when plate and cathode are pulsed simultaneously. The circuit shown was so arranged that there were 400 volts on the plate and, in addition, a pulse voltage of +1200 d-c volts during the pulse interval. The transmitting efficiency could be improved by using it as a transmitter only. The receiver functions, however, were favored in the design of this circuit. The frequency stability is better than ±0.5 Mc/sec for plate voltage variations of ±10 per cent and temperature variations from −40° to +60°C. The difference between the transmitted and received frequency is less than 1 Mc/sec. As a superregenerative receiver this unit has a sensitivity of $2 \times 10^{-9}$ watt for beacon triggering.

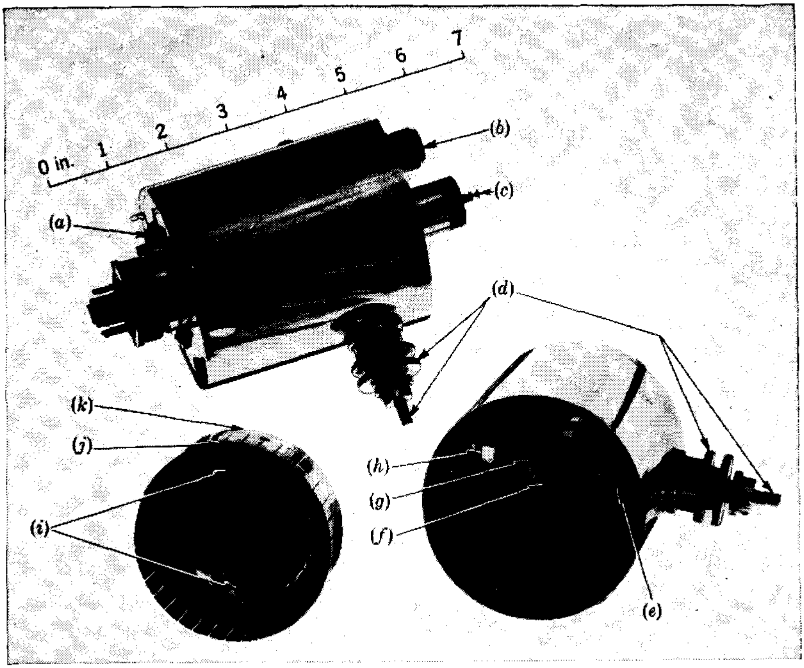


FIG. 14·11.—Photograph of a cavity oscillator for use at 700 Mc/sec. The equivalent circuit is shown in Fig. 14·10 and a cross section in Fig. 14·12. (*a*) Grid lead. (*b*) Antenna connector. (*c*) Plate lead terminal. (*d*) Frequency sweeping mechanism. (*e*) Fixed plate of the tuning capacitor. (*f*) 2C40 plate connectors. (*g*) Plate capacitor. (*h*) Inductive coupling loop. (*i*) Grid-feedback loop. (*j*) Contact fingers. (*k*) Sliding section.

*The* 1000-*Mc/sec Oscillator.*—The cavity oscillator shown in Figs. 14·13, 14·14, and 14·15 is intended for operation in the region of 1000 Mc/sec. It is designed around the small 2C40 lighthouse tube. This cavity may be referred to as a "grid-plate cavity," and is similar to a tuned-plate-tuned-grid circuit. In this case, the grid is connected to the cavity for

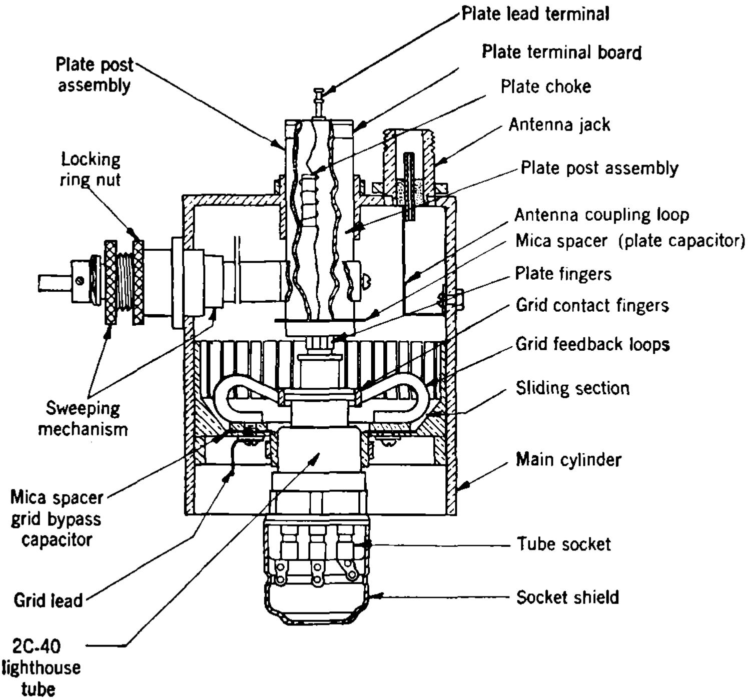


FIG. 14·12.—Cross-sectional view of a 700-Mc/sec cavity oscillator. The equivalent circuit diagram and photograph of this oscillator are shown in Figs. 14·10 and 14·11, respectively.

radio frequency, but not for direct current, since quench voltage is applied to the grid during the reception period. The inner conductor is insulated from the outside shell of the cavity and is connected to the plate. Excitation is obtained by means of cathode fingers, which excite a second cavity. The r-f energy is taken from the cathode cavity by means of an inductive loop. The oscillator is tuned by means of a capacitor on the end of the plate line. The rough frequency adjustment is by means of a slotted rotor plate, which is moved longitudinally so as to give a larger or smaller effective plate area. The fine

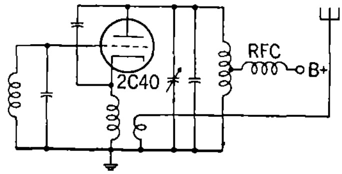


FIG. 14·13.—Equivalent circuit for the 1000 Mc/sec-cavity oscillator shown in Figs. 14·14 and 14·15.

frequency adjustment is made by means of a screw assembly that also increases or decreases the plate area.

This beacon, like that just described, was designed to sweep a band of frequencies. The sweeping of the frequency band is accomplished by means of a motor drive on the variable capacitor mentioned above.

As a transmitter, this oscillator delivers about 70 watts of pulse power to the antenna, with an efficiency of about 40 per cent. The circuit was arranged so that there were 400 volts d-c on the plate and, in addition, a pulse voltage of +1200 volts during the pulse interval. The center frequency could be set anywhere in a 25 per cent band, about which the motor-driven capacitor swept the frequency over a 5 per cent band. The frequency stability is better than ±0.5 Mc/sec for plate voltage variations of ±10 per cent and temperature variations from $-40°$ to $+60°$C. The difference between the transmitted and the received frequency is less than 1 Mc/sec. As a superregenerative receiver this unit has a sensitivity of $2 \times 10^{-9}$ watt.

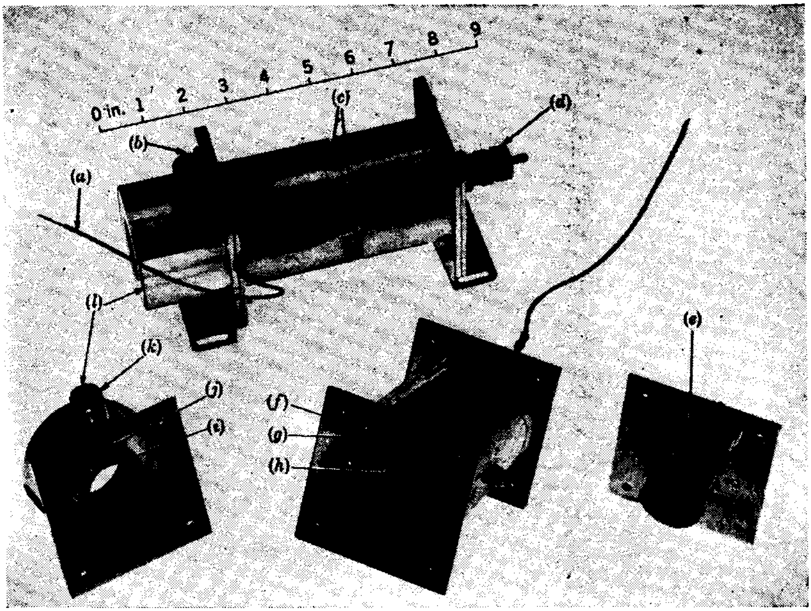


FIG. 14·14.—Photograph of a cavity oscillator for use at 1000 Mc/sec. The equivalent circuit is shown in Fig. 14·13 and the cross section in Fig. 14·15. (a) Grid lead. (b) Antenna connector. (c) Center-conductor supporting screws. (d) Frequency-sweeping mechanism. (e) Rotor of the tuning capacitor. (f) Plate choke. (g) Bypass capacitor. (h) Stator of tuning capacitor. (i) Cathode fingers. (j) Inductive coupling loop. (k) Antenna connector. (l) Cathode cavity.

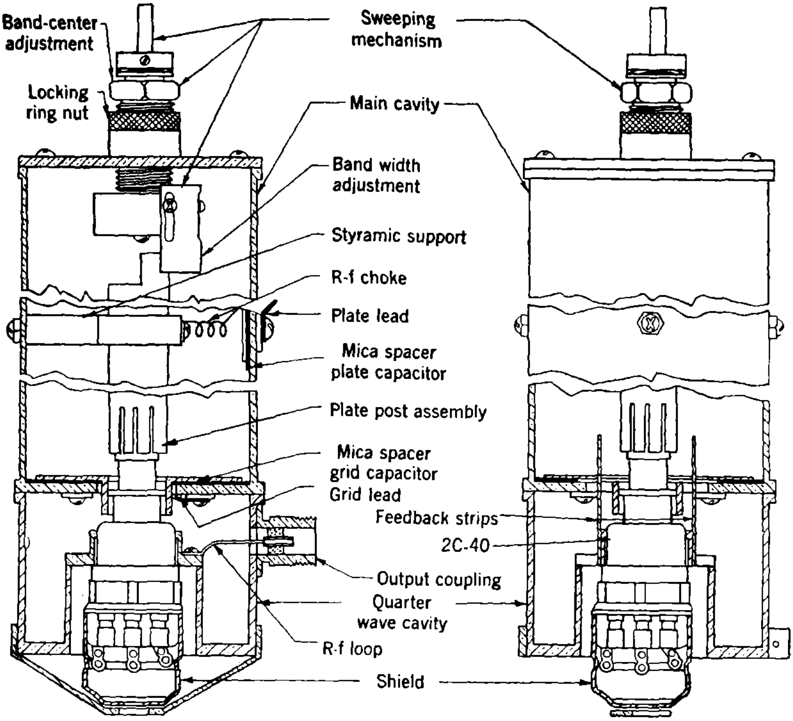


FIG. 14·15.—Cross-sectional view of a 1000-Mc/sec cavity oscillator. The equivalent circuit diagram and photograph of this oscillator are shown in Figs. 14·13 and 14·14 respectively.

## MICROWAVE TRIODE OSCILLATORS

**14·12. Microwave Cavities and Tubes.**[1]—A cavity oscillator for use with triode tubes in the microwave region near 10 cm may have a cavity for the grid-plate region, a second cavity for the grid-cathode region, and a method for providing feedback between the two. If the cavities are separate, the feedback may be supplied by the cathode-plate or grid-plate capacitance of the tube, by coupling loops or probes, or by a matched transmission line coupling the two cavities. In the case of a reentrant cavity, there is no clear distinction between the cavities and the feedback.

The microwave cavity oscillators to be discussed are versions of the reentrant cavity. The exact form of the cavity depends somewhat on the design of the tube used. There are two types of reentrant cavities in general use. One uses a tuned plate line coupled to a coaxial grid line.

---

[1] Secs. 14·12 to 14·15 by K. R. More, J. C. Reed, and P. A. de Paolo.

The other type uses a tuned cathode line and a coaxial grid line. In each case the two coaxial lines are built into the same cavity. Cross-sectional diagrams of the two cavity types are shown in Figs. 14·16 and

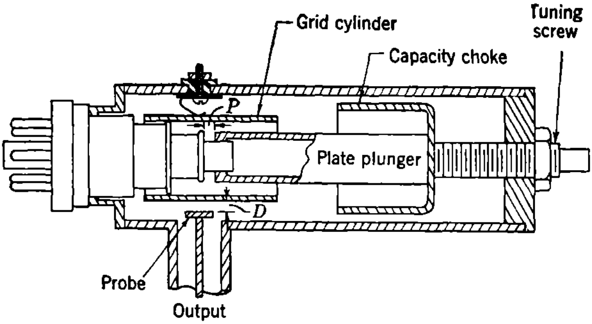


FIG. 14·16.—Cross section of a 10-cm plate-tuned reentrant cavity oscillator, using a 2C40 lighthouse tube. This oscillator will cover the range from 2700 to 3300 Mc/sec with different grid cylinders.

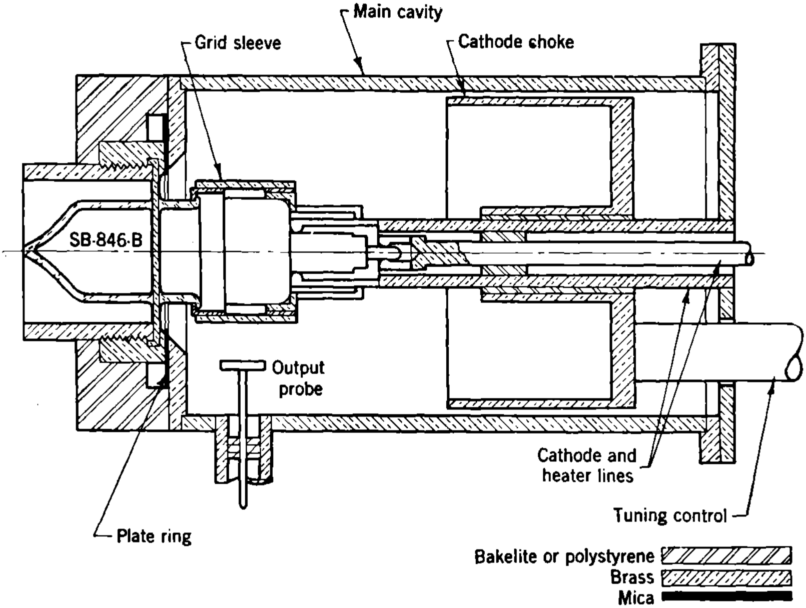


FIG. 14·17.—Cross section of cathode-tuned reentrant cavity.

14·17. The choice of cavity type depends on the tube to be used. The plate-tuned type of cavity is convenient for use with triodes of the lighthouse type, such as the 2C40 and 2C43. Cathode-tuned cavities, on the other hand, are convenient for use with tubes like the 2C36 and SB-846B.

This discussion of microwave cavity oscillators is divided into three parts. The first takes up the tube design considerations made necessary by beacon requirements. The second discusses the cavity design and operational characteristics for the plate-tuned reentrant cavities used with lighthouse tubes. The third part takes up the cathode-tuned reentrant cavity.

**14·13. Tube Requirements.**—The two main requirements placed on microwave oscillator tubes used as transmitters in radar beacons are frequency stability and the ability to give the desired power. Frequency stability is important because a beacon is expected to reply on a given frequency under all conditions of interrogation. The low power-output requirement—usually 50 to 150 watts pulse power—makes impossible the use of some tubes that are satisfactory at higher pulse powers. Higher pulse powers at microwave frequencies are usually obtained from magnetrons.

Three important causes of frequency change are changes in interrogation rate, heater voltage, and input-pulse voltage. Changes in interrogation rate cannot be controlled because they depend on the repetition rate of the interrogator and on the number of radar sets interrogating the beacon. Changes in input voltages sometimes can be reduced by regulators, but it is frequently impossible to hold voltages to better than ±5 per cent. These fluctuations lead to changes in the temperatures of the various tube elements. The consequent expansions and contractions change the interelectrode capacitances and thus the operating frequency.

The effort to reduce frequency drifts affects the designs of both tube and cavity. Both tube and cavity are designed to provide good thermal contact between the plate of the tube and the plate connector of the cavity. This is more difficult in the case of the plate-tuned cavity and lighthouse tube, in which the plate rod makes a sliding contact with the tube plate cap, than it is with the cathode-tuned cavity and tubes like the 2C36 which have a plate flange convenient for fixed contact.

The design of the grid structure is also influenced by stability requirements. Lighthouse tubes now use stretched or "taut" grids to prevent buckling of the grids when the temperature rises. Data on the attainable degree of frequency stability are given in Sec. 14·15, since the designs of both tube and cavity influence frequency stability.

The power-output requirements and the fact that efficiencies of 15 per cent or greater are desirable place limits on the range of power inputs to be used. Such limits may be given roughly as 300 to 1000 watts. Since these cavity oscillators have circuit impedances of 1000 to 2000 ohms, the pulse voltages required are about 500 to 1400 volts.

Experience shows that the 2C43, the only lighthouse triode that carries specifications for pulsed operation, will not operate satisfactorily below about 1800 volts. The effect of a gradual reduction in voltage is a gradual increase in the starting time of the tube, until the starting time exceeds the width of the voltage pulse. It is therefore necessary to use voltages high enough to keep the starting time small—preferably less than 0.1 μsec. Most 2C40 tubes operate very well at pulse voltages of 800 or higher. The 2C40 is not designed for pulsed operation, however, and some tubes give poor performance. A new lighthouse tube, the 2C49, for low-power pulsed operation, was in process of development in 1945.

Tubes of the 2C36 type work well in a cathode-tuned cavity with input voltages as low as 600 volts. The operating frequency is relatively independent of heater voltage.

## REENTRANT CAVITIES

**14·14. Plate-tuned Reentrant Cavities.**—The general features of the plate-tuned reentrant cavity are shown in Fig. 14·16, Sec. 14·12. A detailed photograph of an actual cavity is shown in Fig. 14·18. The design shown in Fig. 14·18 provides for convenient changes of tubes and grid cylinders. The position of the plate rod is controlled by coarse and fine adjusting screws. The capacity-coupled output probe may be moved in and out to maximize the power output. The position of the choke can be changed as required, but such a change involves partial disassembly of the plate-tuning mechanism.

*Frequency.*—The frequency at which such a cavity oscillates is determined mainly by the length of the grid cylinder and the setting of the plate rod. The former is the more important. The frequency is affected only slightly by the position of the terminating choke. As an example, the cavity illustrated in Fig. 14·18, using a 2C40 tube, may be tuned over a range of about 1 cm in the 10-cm region by adjusting the plate rod, for any given grid-cylinder length. On the other hand, for a given plate-rod adjustment, change n grid cylinder length from 0.96 to 1.12 in. will change the wavelength from about 9 to 10 cm. Longer wavelengths can be achieved by the use of even longer grid cylinders and cavities. For beacon service at a specific frequency, the grid cylinder length is chosen to give that frequency when the plate rod is set to the center of its range of motion and a tube of average plate-grid capacity is used. Tubes of other capacities are tuned to frequency by adjusting the plate rod.

*Chokes.*—Two types of terminating choke are illustrated in Figs. 14·16, 14·18, and 14·19. The first figure shows a quarter-wave capacitance choke making no contact with the outer conductor. The second and third figures show a choke making a sliding contact with the outer conductor; it serves as a capacity quarter-wave choke at the plate rod. A polystyrene or polyglas sleeve insulates the plunger from the plate rod. This type of choke has two disadvantages common to all systems involving adjustable contacts: wearing of the contacts and arcing if they are poor. Chokes of both kinds can be clamped at different places along the plate rod to provide for the setting of the choke for optimum power output for each grid cylinder length used. For a system designed for a narrow frequency range only, however, chokes of the capacity type are generally soldered to the plate rod to eliminate sliding contacts.

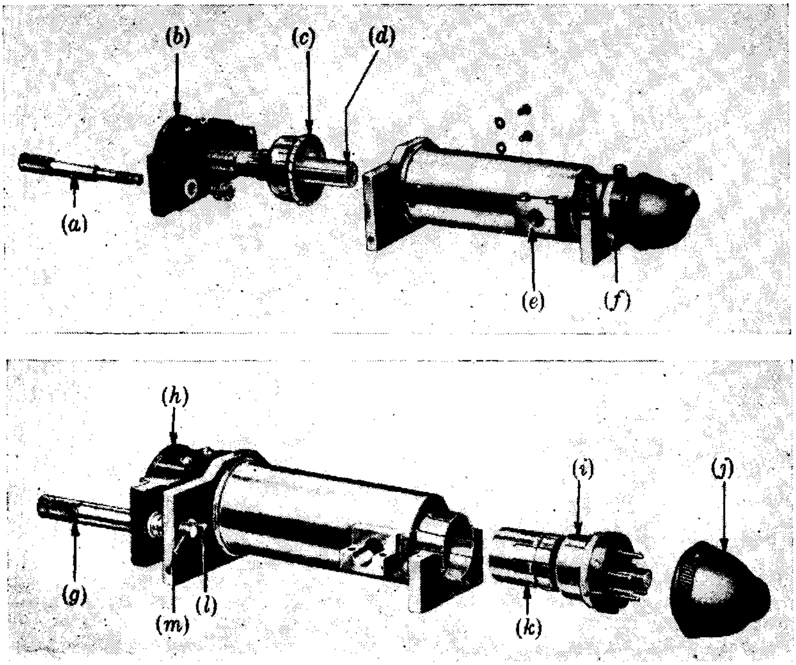


FIG. 14·18.—Exploded view of 2C40 cavity oscillator at 2900 Mc/sec. (*a*) Fine-tuning control. (*b*) Coarse-tuning control. (*c*) Plate choke. (*d*) Plate-tuning plunger. (*e*) Capacitive r-f pickup probe clamp. (*f*) JAN-2C40. (*g*) Fine-tuning control. (*h*) Coarse-tuning control. (*i*) JAN-2C40. (*j*) Tube socket. (*k*) Grid cylinder (in place). (*l*) Stop. (*m*) Limit pin.

More elaborate chokes lead to improved performance, by reduction of leakage and the effect of reflections at the end of the cavity. The chokes discussed above have attenuations of about 15 db over the 9- to 11-cm band. Attenuation as small as this permits enough leakage power to escape to cause interference with beacon receivers that are mounted in the same box as the transmitter. Furthermore, power reflected back into the cavity impairs the oscillator performance for certain critical distances from the choke to the open end of the cavity. Both troubles can be reduced when necessary by the use of a polyiron termination.[1] In the design of more elaborate chokes, it is necessary to keep the capacitance from the plate rod to ground to a reasonably low figure. Otherwise the capacitance will load the modulator heavily, and the voltage pulse will be distorted as a result. In spite of the disadvantages of the chokes shown in Figs. 14·16 and 14·18, they are used widely in pulsed transmitters because of their simplicity.

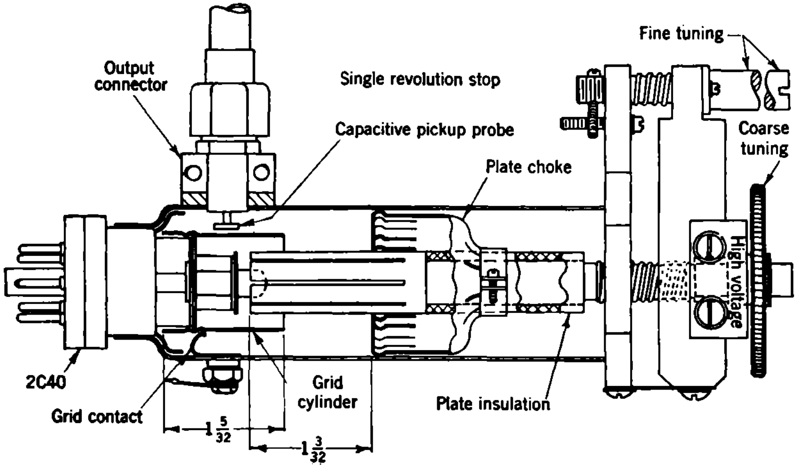


FIG. 14·19.—Cross section of a plate-tuned reentrant cavity oscillator, using a 2C40 lighthouse tube. Note that the plate choke makes a sliding contact with the outer conductor. Dimensions given are for operation at frequencies between 2700 and 3000 Mc/sec.

*Output Coupling.*—The size of the output probe button influences the maximum power output attainable and the ease of adjusting the probe to give maximum power output. In general, a large probe gives greater power output and a broader power maximum than a small probe. A probe diameter of 0.25 to 0.31 in. is satisfactory for the cavity shown in Figs. 14·18 and 14·19. The larger diameter broadens the maximum of the power vs. probe-position curve but does not raise it.

*Grid Connection.*—The grid contacts are insulated from the cavity shell. This permits operation with a grounded grid or with a bias resistor between grid and ground. These cavities give satisfactory performance when one grid contact only is used; however, the use of three contacts symmetrically located around the outer tube of the cavity is advisable

---

[1] See, e.g., *Technique of Microwave Measurements*, Vol. 11, Chap. 12, Radiation Laboratory Series.

for both mechanical and electrical reasons. Vibration may result in poor contact if one contact only is used and the increase in power output gained by using a grid resistor is greater with three contacts than with one.

The output probe and grid contacts lie in a plane perpendicular to the axis of the cavity. The location of this plane is determined experimentally; it should be one that gives suppression of oscillations at unwanted lower frequencies.

*Construction.*—The materials used in cavity construction may be chosen to compensate for the frequency shifts that accompany changes in the temperature of the tube and cavity. One such method is the use of a steel outer shell to reduce the over-all cavity expansion and of a plate rod having a high coefficient of expansion to compensate for the expansion of the tube and the outer shell. The plate contacts must be made of spring material, like beryllium copper or phosphor bronze, in order to provide a good sliding contact. A section of duraluminum for the outer end of the plate rod increases the expansion.

The insulating materials used for mounting the plate rod and insulating the choke from it must not distort under pressure at the highest temperatures likely to be reached during the operation of the beacon. High-quality linen-base bakelite is satisfactory for the tuning arm. Polyglas-D molded to the plate rod insulates the choke. Polystyrene or copolystyrene sleeves are not satisfactory because the distortions to which they are subject even at ordinary operating temperatures result in frequency shifts of the order of 3 Mc/sec.

**14·15. Operational Characteristics.**—The position of the terminating choke has an important effect on the operation of the lighthouse-tube reentrant cavity oscillator. Power output, input current, optimum output probe position, and wavelength are plotted in Fig. 14·20 as functions of the distance *d* of the choke from the end of the plate rod, under conditions of constant input voltage, grid-cylinder length, and plate-rod position. These curves show that the position of the choke must be held within reasonably close limits to keep the power output near the maximum value. Furthermore, since the input current increases with *d*, it is advisable to keep *d* slightly below the value giving maximum power output. This is particularly true if a line-type pulse generator is used to supply the power. The low impedances associated with large values of *d* load the generator excessively, and a lower pulse voltage results. As has already been mentioned, too low a voltage may result in poor performance of the tube.

The frequency of oscillation of any tube in a given cavity is determined by the grid-to-plate capacitance. This effect is shown in Fig. 14·21. The over-all scatter of wavelength in this particular cavity due

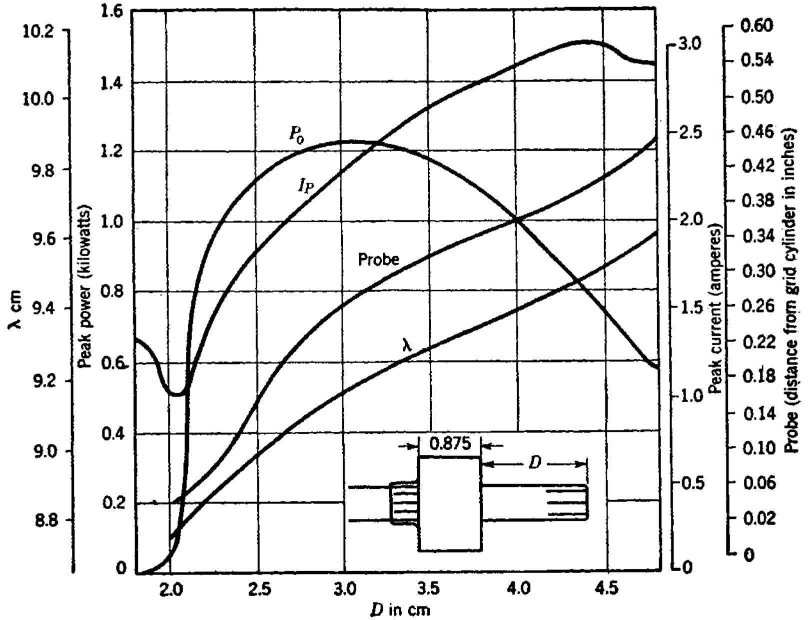


FIG. 14·20.—The effect of plate-choke position on wavelength, pulse-power output, input current, and optimum probe distance, in the 10-cm reentrant cavity oscillator. Data are for the average of several 2C43 tubes.

to the permissible variation in $C_{gp}$ is about 0.5 cm. This reduces the usable band of wavelengths from the 1.1-cm range provided by the plate tuning to a usable range of 0.6 cm, for tubes whose grid-plate capacitances differ most from the mean value. The additional scatter due to cavity tolerances reduces the usable range of wavelengths for a given grid-cylinder length and choke position to about 0.5 cm. The use of an extension on the plate cap extends the tuning range.

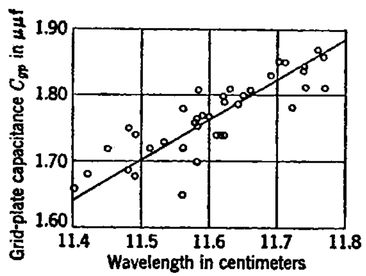


FIG. 14·21.—Wavelength as a function of grid-plate capacitance for a fixed setting P of the plate rod (as shown in Fig. 14·14). Data are for 2C43 tubes in the same cavity, under identical operating conditions.

The performance chart of Fig. 14·22 shows the behavior of a typical reentrant cavity oscillator. Contours of constant grid-bias resistance and of constant power output are drawn on a current vs. plate-voltage diagram. The use of grid-bias resistors of approximately 50 to 150 ohms is useful as a means of increasing the impedance. Values as high as 500 ohms impair performance at the low-voltage limit of operation.

The manner in which the output power varies with the position of the output probe depends on the position of the choke and on the adjustment of the plate rod. The power may fall off from a maximum for the probe touching the grid cylinder, as shown in Fig. 14·23*b*, or it may drop to a minimum and rise to another maximum before falling off, as in Fig. 14·23*a*. In either case, it is advisable to set the probe at a position enough beyond that which gives the maximum so that the power drops to about 90 per cent of the maximum. A Rieke diagram (see Sec. 13·2) for the oscillator shows that the pulling figure (see Sec. 13·3) is appreciably lower when the probe is set in this way than when it is set for maximum power.

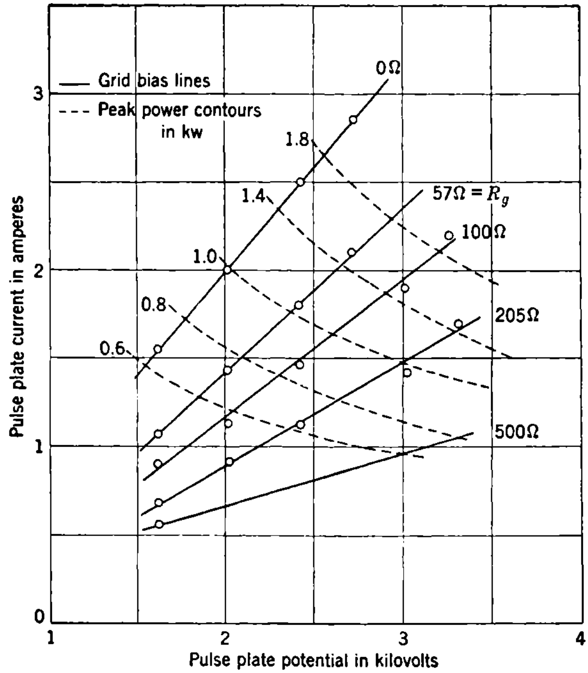


FIG. 14·22.—Average performance chart of 10 2C43 tubes in a plate-tuned reentrant cavity at about 11 cm, as a function of grid-bias resistance.

The examples used above to illustrate operational characteristics are all based on the 2C43 lighthouse tube, even though this tube is used only at power levels higher than are needed for lightweight beacons. More complete data are available for the 2C43 tube, however, than for the

2C40 tube, which is used in some beacons. Data available for other tubes used in beacons agree in general behavior with data on the characteristics of the 2C43. Voltages, currents, and power figures, of course, are lower for the 2C40. Some data on power output and efficiency for 2C40 tubes used in cavities of the type shown in Fig. 14·18 are given in Table 14·1.

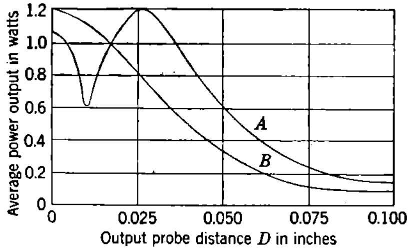


FIG. 14·23.—Power output vs. position of output probe for the plate-tuned reentrant cavity shown in Fig. 14·14. Curves are for two different settings of plate, choke, and plate rod.

TABLE 14·1.—WAVELENGTH, INPUT PULSE VOLTAGE, PULSE CURRENT, INPUT POWER, OUTPUT POWER AND EFFICIENCY, FOR TWO 2C40 TUBES*

| Tube | $\lambda$ (cm) | $E_p$ (volts) | $I_p$ (amp) | $P_i$ (watts) | $P_o$ (watts) | Eff (per cent) |
|---|---|---|---|---|---|---|
| *A* | 9.85 | 987 | 0.57 | 560 | 65 | 12 |
| | 10.24 | 940 | 0.58 | 535 | 115 | 21 |
| | 10.90 | 917 | 0.61 | 560 | 113 | 20 |
| *A* | 9.85 | 1078 | 0.62 | 670 | 97 | 15 |
| | 10.24 | 1036 | 0.63 | 650 | 199 | 31 |
| | 10.90 | 987 | 0.65 | 640 | 183 | 29 |
| *B* | 9.81 | 910 | 0.57 | 520 | 122 | 23 |
| | 10.23 | 889 | 0.62 | 550 | 155 | 28 |
| | 10.97 | 855 | 0.63 | 540 | 148 | 27 |
| *B* | 9.81 | 994 | 0.63 | 625 | 164 | 26 |
| | 10.23 | 980 | 0.64 | 625 | 204 | 33 |
| | 10.97 | 932 | 0.66 | 615 | 184 | 30 |

* Wavelengths are for the low limit, middle, and high limit of the tuning range obtained by plate-rod adjustments. Data are for two settings of input pulse.

*Frequency Stability.*—The change in frequency of a 2C40 10-cm oscillator, resulting from a change in filament voltage from 6.3 to 6.9 volts, is from 0.5 to 3.0 Mc/sec, averaging about 2.0 Mc/sec. The change resulting from a pulse-repetition-frequency change from 500 to 2000 cps is about 1 Mc/sec, varying from 0.3 to 1.5 Mc/sec. The frequency change accompanying an input voltage change of 100 volts is about 1 Mc/sec. Accordingly, it is desirable to stabilize the heater current and the applied pulse voltage for maximum frequency stability.

**14·16. Cathode-tuned Reentrant Cavities.**[1] *Cavity Design.*—The grid element of the cathode-tuned reentrant cavity is decisive in determining the oscillator frequency. Changing the position of the terminating choke will shift the frequency less than 1 per cent and the efficiency will vary considerably over this tuning range. This feature complicates the tuning mechanism, but the oscillator is, in general, more stable than the plate-tuned type of reentrant oscillator.

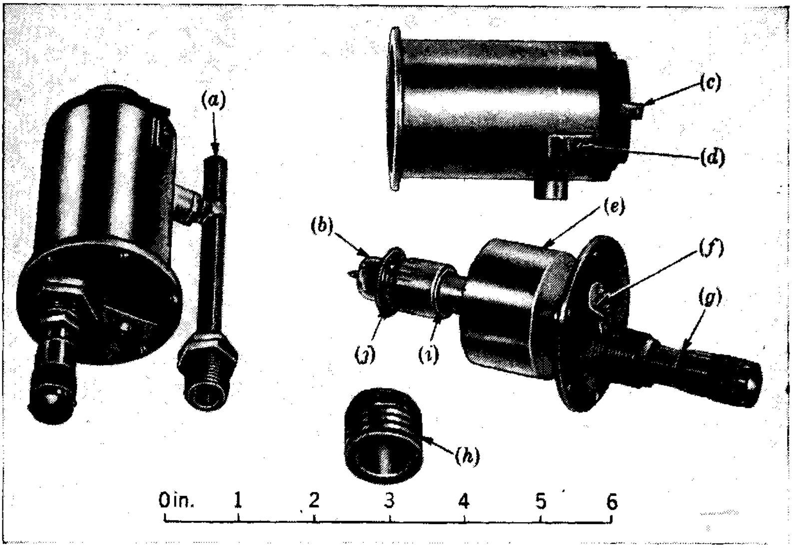


FIG. 14·24.—Cathode-tuned reentrant cavity 10-cm oscillator using the SB-846B triode. A diagram is given in Fig. 14·17. (*a*) Transmission line. (*b*) SB-846B tube. (*c*) Insulated plate connection. (*d*) Grid-ground connection. (*e*) Cathode choke. (*f*) Heater connection. (*g*) Micrometer tuning control. (*h*) Tube-retainer plug. (*i*) Grid sleeve. (*j*) Plate flange. (*k*) Output connector.

There are two methods of tuning the cavity. One is changing its length, and the other is capacitance-loading of the grid element. When either of these methods is used, the cathode cavity must be retuned to realize maximum efficiency. The cavity shown in Figs. 14·17 (Sec. 14·12) and 14·24, using the SB-846B tube, is tuned by varying the capacitance between the end of the grid element and the inner conductor of the cathode cavity. Under these conditions, changing the position of the

---

[1] By J. C. Reed.

cathode-cavity terminating choke will also tune the grid element. If this system is used, a 10 per cent tuning range can be realized, with a difference of power output over the tuning range of less than 20 per cent.

The terminating choke does not make contact with either the inner conductor or the outer conductor. It obtains its mechanical support by sliding on a polystyrene sleeve that surrounds the inner conductor. The position of the choke is controlled by means of a screw adjustment.

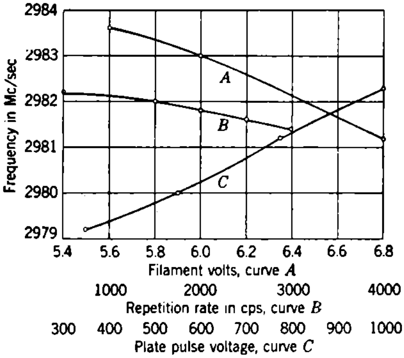


FIG. 14·25.—The frequency-stability characteristics of the oscillator of Figs. 14·17 and 14·24. (a) Frequency vs. repetition rate, with filament-voltage constant at 6.3 volts and plate pulse at 800 volts. (b) Frequency vs. filament voltage, with pulse-repetition frequency constant at 800 cps and plate pulse at 800 volts. (c) Frequency vs. pulse plate voltage, with filament-voltage constant at 6.3 volts and pulse-repetition frequency at 800 cps.

*Operating Characteristics.*—The efficiency of the SB-846B tube is 15 per cent or better over the tuning range 3100 to 2900 Mc/sec when used in the cavity shown in Fig. 14·24. At $E_f = 6.3$ volts, the rated heater voltage, and $E_p = 800$ volts, the efficiencies of 10 sample tubes were between 20 and 28 per cent. At $E_f = 5.8$ volts, $E_p = 800$ volts the efficiencies of the same tubes were between 17 and 27 per cent. The pulse power output was approximately 100 watts. The excellent frequency stability of the oscillator is shown in Fig. 14·25.

# CHAPTER 15

# POWER SUPPLIES AND PERFORMANCE TESTING

By B. W. Pike, M. J. Cohen, and J. J. G. McCue

## POWER SUPPLIES

By B. W. Pike

Radar beacons have, in general, the same power-supply requirements as other complex electronic equipment that must be reliable: a suitable primary power source and suitable heater, plate, and bias power supplies for electron tubes. In addition, a few special power-supply problems are met in beacons. These arise from the use of magnetrons, planar triodes, and velocity-modulated tubes, and from the need for a high order of frequency stability of those tubes used as self-excited oscillators. Only points of special interest to beacon designers will be discussed in this chapter. General theory and design data on power supplies are well covered in generally available texts, handbooks, and periodicals.

*Beacon Power-supply Requirements.*—The particular requirements of beacon power supplies, in addition to the basic requirements of long life and a high degree of reliability, are listed below.

1. Good voltage regulation over a wide range of loads, primary supply voltage, and temperature.
2. Power-control circuits suitable for automatic, unattended operation of beacons used as continuous-duty navigation aids.
3. Both high no-load efficiency and high full-load efficiency, because of the intermittent nature of the load.
4. Freedom both from self-generation of noise and from pickup of extraneous electrical noise that would trigger the beacon because of the limited duty ratio of the beacon.

*Primary Power Sources and Power Supplies.*—Primary sources of electric power of three types have been used for radar beacons: (1) commercial a-c power; (2) the alternator driven by a gasoline engine; and (3) the storage battery, used either alone or with a generator or other charger connected to it.

Three types of power supply are used widely in beacons. Beacons designed to operate from a-c power lines use single-phase rectifier supplies.

Beacons intended for portable ground operation often use a storage battery to feed a vibrator supply. Some battery operated ground beacons are supplied from dynamotors. Almost all airborne beacons have a dynamotor with a carbon-pile primary voltage-regulator. A few airborne beacons operate from variable-frequency constant-voltage alternators, driven either by the engine or by a d-c motor.

**15·1. Prime Power Sources.** *Commercial A-c Power Lines.*—Because even the largest beacon yet made requires no more than 2 kva, single-phase alternating current has been used exclusively in a-c operated beacons. Recently designed large beacons can be operated either at 115 or at 230 volts ± 7 per cent, at 50 to 70 cps; smaller beacons, at 115 or 230 volts ± 10 per cent, 50 to 2400 cps.

In some early beacons, an a-c voltage regulator was needed even on well-regulated power lines. The resonant type of constant-voltage transformer was used. This type of regulator is satisfactory provided certain factors are taken into account. The first is that these transformers usually have a large third-harmonic component in their output voltage. This means that the high-voltage supplies, if designed for a pure sine wave, may not deliver enough voltage. Second, these regulators are frequency sensitive and, therefore, unsuitable except for constant-frequency lines. Third, they usually employ electrolytic capacitors and therefore may not be satisfactory over a wide range of temperatures. It is preferable to make the entire beacon capable of accommodating the expected variations of line voltage by using regulated power supplies and constant-current modulators (see Sec. 12·8).

*Engine-driven Alternators.*—Small alternators driven by gasoline engines are widely used for operating a-c beacons and battery chargers in locations where a-c power is not available. Except in the large sizes (5 kw and up), these power supplies often suffer from relatively poor voltage regulation and from the need of frequent removal of carbon from the engine. Furthermore, the small engines are often unreliable. Therefore, for a-c beacons that must maintain continuous operation, a dual installation is necessary.

For such beacons, engine generators like the Army PE-214B (see Fig. 15·1) may be used successfully. The PE-214B weighs 50 lb and produces 300 va at 115 volts, 60 cps. This unit comprises a single-cylinder, two-cycle gasoline engine, direct-coupled to a permanent-magnet alternator. Since it uses a voltage-controlled throttle rather than a speed governor, the regulation is satisfactory.

Inasmuch as power supplies can be made insensitive to variations of frequency, a voltage-regulated variable-frequency alternator coupled to the engine of an aircraft provides an excellent prime power source for airborne electronic equipment. Audio-frequency alternating current

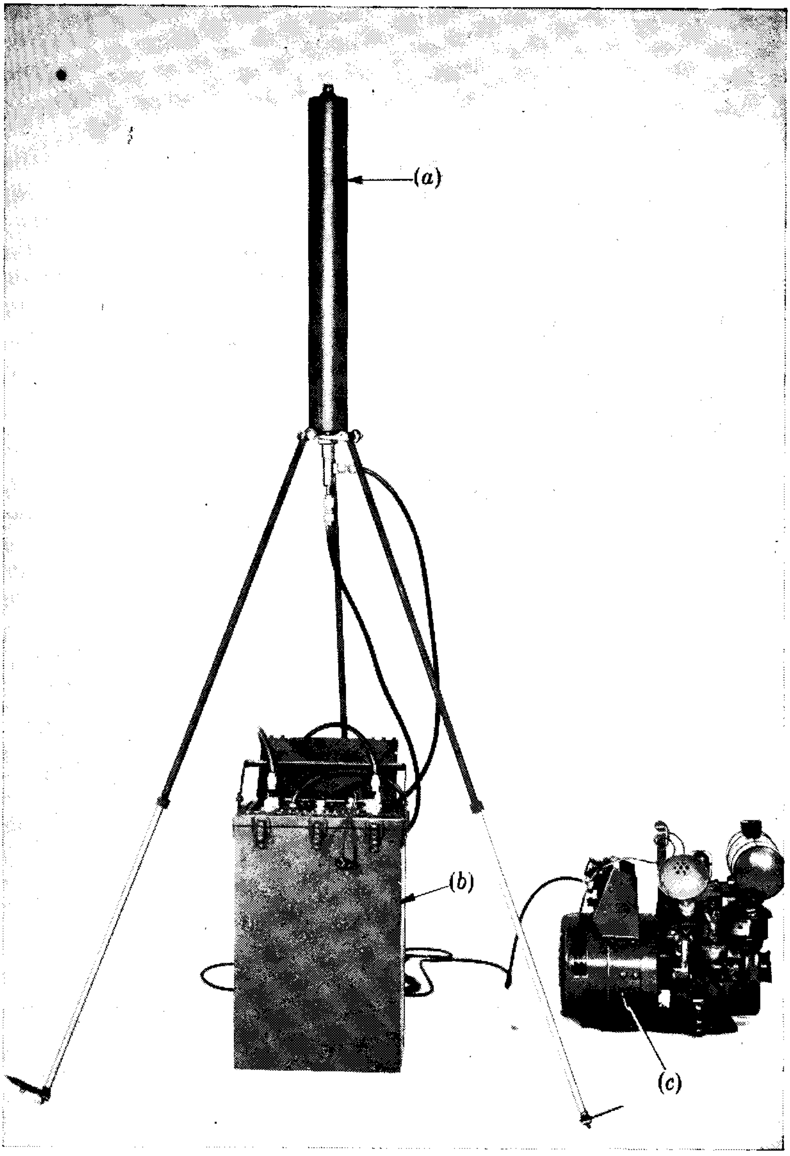


FIG. 15·1.—A lightweight 10-cm a-c–operated beacon (AN/UPN-2) with an engine-driven alternator prime power source. (a) Antenna. (b) Transmitter-receiver. (c) PE214B engine-driven alternator.

appears to be the best primary electric power[1] for reliability, light weight, freedom from radio interference, and the like.

*Dry-cell and Storage-cell Batteries.*—Because of their relatively poor regulation, failure at low temperatures, and poor storage properties, dry cells have not been used for beacons. Except for the British lightweight 220-Mc/sec Eureka beacon, alkaline storage cells have not been used widely in beacons because of their poor low-temperature capacity, unfavorable discharge-curve shape, and unavailability during the war. Instead, the lead-acid type of cell, readily available in several "spill-proof" forms, with a relatively flat discharge curve, was generally used. For applications in which the battery can be kept sufficiently warm, however, the alkaline battery is advantageous because of its high capacity and ruggedness.

The batteries used for beacons have varied from a single 2-volt cell to the 28-volt battery system used in military aircraft. Several beacons were equipped with auxiliary battery chargers, both for charging batteries and for "floating" the batteries for continuous operation of the beacon when prime a-c power was available. Such chargers generally consist of a transformer for operation from 115 to 230 volts, 50 to 2400 cps, with a tapped secondary for supplying a dry-disk rectifier of the selenium type.[2,3] Wind-driven battery-charging generators are useful auxiliaries in remote permanent stations.

**15·2. Alternating-current Power Supplies.**—High-performance power supplies of this type are the most flexible, simple, and reliable of all those used in beacons. Among them four classes can be distinguished: high-voltage supplies for magnetron modulators; medium-voltage supplies for grid-pulsed oscillators, modulator drivers, and so on; low-voltage supplies for receivers, coders, and so on; and low-voltage grid bias supplies. Medium, low-voltage, and bias supplies are often combined into multiple-output units. Examples will be described in Secs. 15·8 and 15·9.

*High-voltage Supplies.*—The high-voltage a-c power packs for magnetron modulators fall in the range 15 kv at 40 ma to 3 kv at 5 ma, according to the type of modulator and magnetron. As was shown in Sec. 12·8, a magnetron with a triode switch-tube requires very good regulation of the power-supply voltage to maintain the required frequency

---

[1] For these reasons, the MIT Radiation Laboratory developed several versions of small engine-driven high-frequency alternators for lightweight radar and beacon equipment. These involved rotary steam engines, turbines, etc. See *Components Handbook*, Vol. 17, Sec. 12·3, Radiation Laboratory Series.

[2] J. E. Yarmack, "Selenium Rectifiers and Their Design," *Elec. Eng.*, **61**, 488 (1942).

[3] *Reference Data for Radio Engineers*, Federal Telephone and Radio Corp., New York, 1943.

stability. Triode switch-tubes are no longer used for magnetron transmitters in beacons; tetrodes are used exclusively because no regulation of the high-voltage supply is required.

*Medium-voltage Supplies.*—Medium-voltage a-c power supplies for drivers, for the screens of constant-current tetrode modulators, and for grid-pulsed oscillators fall within the range 500 to 1500 volts, 5 to 25 ma. For modulator drivers, the voltage need not be electronically regulated because a driver stage can be designed to deliver a small excess of modulator-grid drive at the lowest expected plate voltage. Higher plate voltage is accommodated by driving the modulator grid positive through a series grid resistor.

Both for grid-pulsed oscillators and for the screen grids of constant-current modulators, however, it is essential to have good electronic regulation. In drivers for constant-current modulators, it is common to use the supply voltage of the driver with an electronic regulator for supplying the screen voltage of the modulator. This voltage must be adjustable in order to control the current level of the constant-current modulator.

*Low-voltage Supplies.*—Low-voltage a-c supplies for receivers, coders, and gas-filled tube modulators are those in the range from 100 to 400 volts and 10 to 200 ma. These supplies are always regulated either with a hard-tube or gas-filled-tube regulator, except when used for certain noncritical stages of receivers. Because of the filtering action of these regulators, only a small value of filter capacitance is necessary. It is not difficult to reduce ripple to 0.05 volt without the use of electrolytic (high-capacity) filter condensers. Mineral-oil-filled condensers are preferable to those filled with the relatively temperature-sensitive polar dielectric fluids.

An electronically regulated power supply, because of its very low generator impedance, is well suited to supplying two different loads which might interact through a common coupling. For example, a receiver and a coder have to be powered from separate supplies, or at least from separate filters, unless regulation is used.

*Bias Supplies.*—Negative bias supplies for a-c beacons differ little from low-voltage supplies, except that the load current is small. For circuits that draw varying grid current, a low-impedance bias source is necessary to prevent "grid-leak" action. Unless integral multiples of the voltages available from VR tubes can be used, it is necessary to use a hard-tube regulator which can be adjusted to the desired voltage for this type of bias load. It is usually possible to use either no regulation or VR-tube regulation for beacon bias supplies. For simplicity it is sometimes desirable to have a dual low-voltage and bias supply by using a single transformer-rectifier to feed a VR-tube voltage divider with the desired point grounded. Alternatively, separate rectifiers can be fed from the same transformer voltage.

*Rectifiers.*—All high-voltage a-c beacon supplies have used vacuum-tube rectifiers. Vapor rectifiers have not been used because of their poor temperature characteristics. High-voltage, low-current selenium rectifiers, developed during the war, have proved satisfactory in vibrator supplies (see Sec. 15·3).

Because of the relatively large currents sometimes required from low-voltage supplies, it is desirable to use efficient rectifiers. Gas-filled diodes and selenium dry-disk rectifiers now appear to be satisfactory. The gas-filled rectifiers eliminate the temperature problem, but generate r-f noise that is difficult to filter. For this reason, the noise-free, temperature-insensitive, selenium dry-disk rectifier will probably be preferred.

*Transformers.*—Transformers for beacon use are usually hermetically sealed in metal containers; they are reliable but large and heavy. For transformers of small sizes, the new plastic-coating methods will probably cut down size and weight effectively without sacrificing the waterproof construction. Such transformers, when properly made, give good service.

**15·3. Vibrator Power Packs.**—Only a few special aspects of a-c power supplies were touched on in Sec. 15·2 because a-c supplies have been rather fully discussed in previously published literature. This is not true for vibrator power supplies (discussed, in Sec. 12·7 of *Components Handbook,* Vol. 17, Radiation Laboratory Series). Few authentic data on vibrator power supplies have been published in the United States because, to paraphrase an excellent British report,[1] vibrator manufacturers keep to themselves whatever scientific or empirical approaches to the design of associated circuits they have been able to develop.

For small battery-operated beacons requiring high voltages, the vibrator type of supply is now preferred to rotating machinery. Direct-current generators for high voltage are hard to build to a reasonable size because of the difficulties of commutation at high voltage. An a-c machine to supply an ordinary transformer-rectifier power supply has inherent advantages, but no a-c machines small enough for lightweight beacons are available. For the power levels now possible from vibrators, vibrator supplies are always superior to rotating machinery in efficiency and weight. Field experience has shown, furthermore, that hermetically sealed vibrators, including those of the synchronous rectifying type, are thoroughly reliable.

There is not much difference between dynamotor and vibrator supplies in the amount of radio interference they cause. The dynamotor

---

[1] E. H. Niblett, *Vibrators and Vibratory Converter Techniques in America,* Signals Research and Development Establishment, Report 862, May 1943.

may appear to be simpler, but it can cause increased interference at high altitudes or as the brushes and commutator wear. In any case, the operation of a beacon from a single low-voltage battery requires very careful elimination of noise. (See Chap. 16.)

Vibrator supplies for beacons, like a-c operated supplies, fall into four classes: high-, medium-, and low-voltage plate supplies, and negative bias supplies.[1] High-voltage vibrator supplies, for low-power magnetron modulators, grid-pulsed oscillators, and so on, usually fall in the range 1000 volts at 6 ma to 3000 volts at 4 ma. For these high voltages, the synchronous rectifier is not usable. Vacuum, gas-filled diode, or selenium, rectifiers are necessary. The gas-filled diode and the selenium dry-disk rectifier have the great advantage that it is not necessary to heat rectifier filaments with power taken through the vibrator and transformer.

Miniature vacuum rectifiers, gas-filled rectifiers, and VR tubes, which have become available recently, should simplify the design of small vibrator supplies.[2] With vibrator supplies, the additional radio interference caused by gas-filled rectifiers is negligible. The new gas-filled tubes, therefore, may have some advantages of weight, size, efficiency, and reliability over selenium rectifiers for this type of service.

It is sometimes possible to combine both low- and high-voltage supplies by using a synchronous-rectifier supply for the low voltage and having a separate transformer winding and rectifier for the high voltage. For all but the smallest beacons, however, the sum of the power requirements of the two supplies exceeds the permissible load on vibrators currently available.

For reasons of power economy, high-voltage supplies for battery-operated beacons ordinarily are not regulated. Either constant-current modulators are used or the variations in performance that result from poor regulation are tolerated.

Medium-voltage vibrator supplies are used for the screen grids of low-power constant-current modulators and for the plates of drivers, blocking oscillators, and so on. The voltages required range from 300 to 500 volts at 10 ma or less. Because these voltages are too high for synchronous rectifiers, selenium, gas-filled, or vacuum rectifiers have been used. When regulation is required, VR tubes are customarily used. Tubes of small size and low filament power, suitable for electronic regulation, are not available.

In a battery-operated vibrator supply, the range of input voltage is usually large. It can extend from 10.5 to 12.5 volts for a 12-volt lead-acid battery, a range of 20 per cent. This means that, because both VR

---

[1] Examples are given in Sec. 15·10.

[2] See *Components Handbook*, Vol. 17, Chap. 16, Radiation Laboratory Series.

tubes and electronic regulators dissipate power, and because long battery life is desired, every effort should be made to design voltage-compensating circuits that do not require regulation. This has been accomplished for most receivers and for a few other circuits by allowing the bias voltage to increase negatively in proportion to an increase in plate voltage. Critical circuits such as discriminators, range coders, and constant-current modulators, however, require regulation for the plate, screen, and bias voltages.

Since these medium voltages are not used for high-gain circuits like beacon receivers, the filtering of ripple is not critical. Oil-filled condensers of relatively low capacitance, plus resistance, or resistance and VR tubes are sufficient.

Low-voltage vibrator supplies supply voltages in the region near 250 volts at 50 ma. This region is that of the common supply for automobile radios; it makes use of the self-synchronous rectifier type of vibrator. In beacons, these supplies can be used for noncritical circuits in receivers, and so on, and are not regulated. The use of these supplies in beacons presents no new problems except that removal of radio noise may have to be more complete than in such an application as automobile radio. Electrolytic capacitors are practically essential for vibrator supplies to be used with beacon receivers.

Vibrator bias supplies usually are of low power and low voltage, such as 100 volts and 10 ma. They are often combined with either the high-, medium-, or low-voltage supply. This is done either by adding a separate transformer and a secondary rectifier, by using a center-grounded voltage divider, or by using reversed rectifiers on a medium voltage secondary. When a synchronous rectifier is used, a separate bias winding is necessary.

**15·4. Dynamotor Supplies.**—Because of the difficulty of commutating high voltage in machines of size commensurate with the power required for battery-operated radar beacons, only beacons that can operate on less than about 500 volts direct current use dynamotors. This means that, even for low-power beacons, only low-impedance modulators with stepup pulse transformers can be used with dynamotors.

The usual dynamotor supply is designed to work at 12 or 24 volts and to make use of a series carbon-pile regulator which delivers 19.5 volts from a 24-volt supply.[1] These regulated dynamotors have now attained good performance under the conditions of altitude, temperature, and vibration encountered in military aircraft. Although, in the past, dynamotors have been preferred over vibrators for beacons, because they were thought to be more reliable and accessible to inspection, the sealed vibrator now competes successfully with the dynamotor for battery-

---

[1] A typical airborne-beacon dynamotor supply is described in detail in Sec. 15·11.

operated beacons in applications in which more than 500 volts are required.

The problem of eliminating radio noise caused by brush-sparking in dynamotors is almost as difficult as that encountered in vibrators. It is discussed in Chap. 16.

**15·5. Filament Supplies.**—Filament supplies do not offer any great problems in beacons. Magnetrons are usually pulsed negatively on the cathode. The filament transformer, therefore, must have adequate insulation to withstand the high-voltage pulse and very low capacitance to ground. Such special transformers are not difficult to design. When pulse transformers are to be used to drive magnetrons, the filament power can be obtained from the ground end of the secondary through a bifilar secondary winding, thereby allowing the use of an ordinary filament transformer.

Some magnetrons require that the cathode heater be turned off at high duty ratios, the cathode being heated by electrons that return to it. In beacons, the duty ratio varies; a relay in series with the magnetron anode current may be used to turn the filament off at high duty ratios. More satisfactory results can be obtained with saturable core reactors or filament transformers.

Planar triode oscillators at high frequencies change in frequency with change of heater voltage so that regulation is required. In battery-operated beacons, either no regulation is provided or the carbon pile that regulates the dynamotor is used. Beacons operating on alternating current use a constant-current ballast tube of the iron-filament type. Because of variations in commercial tube heaters and ballast tubes, it is usually necessary for optimum regulation to include two adjustable resistors in the circuit. Such a ballast-tube regulator is described in Sec. 15·9 as a part of the lightweight a-c beacon power supply.

**15·6. Miscellaneous Power Supplies.**—A number of types of power supply that might be used in beacons will be discussed below, although they have not been used widely until now (1946).

A power supply using a small rotary converter and rectifier for stepping up d-c voltage was used in some German military beacons, but no attempts to develop this type of supply are known to have been made in the United States. Although airborne beacons are often supplied by converters giving alternating current at high audio frequencies, small converters have not been built into airborne beacons as part of each battery-operated power supply. This type of power supply appears to involve a roundabout power conversion compared to that of the dynamotor or the vibrator, but it has some advantages for small d-c–operated beacons because it combines the reliability of the dynamotor with the flexibility and reliability of the a-c power supply. In addition, it is flexible as to input voltage. A simple exchange of a plug-in converter allows a different battery voltage to be used, and because the power-supply unit can also operate from a-c sources, a-c–d-c operation is possible.

Vacuum-tube oscillator supplies have been considered for low-power requirements. Because of the simplicity and light weight of the high-voltage transformer needed, such power supplies are especially useful when high voltages at very small currents are required. This type of supply has not been used, however, because the high-voltage requirements of small beacons are at a lower impedance range than would justify its use. It appears unlikely that such tube-type inverters can ever compete with vibrators at supply voltages as low as 26 volts because the tube drop is such a large fraction of the available voltage. For higher supply voltages, however, tube inverters can be reasonably efficient and fairly reliable.

**15·7. Power-control Circuits.**—Beacon power-control circuits perform the following functions: overload protection, preheating of cathodes, indication of trouble by alarms, conservation of power between interrogations, and automatic restoration of power following overload or failure of prime power.

*Overload Protection.*—The simplest and most commonly used overload protection is the fuse. For simple beacons, only fuses are used for protection. In more complex beacons, circuit breakers backed up by fuses are preferred. Fuses or thermal circuit breakers are always used in the prime-power lines. In more complex beacons, each power supply has its own primary fuses; the critical circuits, such as the transmitter, have overload relays.

Overload relays are necessary because of the occasional internal spark-over that will occur in pulsed vacuum tubes. The overload relay disconnects the high-voltage supply to extinguish the arc. The usual rapid-acting overload relay is made to reclose the high voltage after a second or less. In the simplest beacons, this overload relay merely continues to reclose if there is a short circuit. If the proper size fuse is placed in the high-voltage supply primary, the fuse will be heated a little more on each overload cycle and will blow after about a dozen successive reclosings. A slow-release slug-type overload relay of this kind is shown in the power supply of the a-c lightweight beacon described in Sec. 15·9 (Fig. 15·7).

In more complicated beacons, like the large ground and shipboard beacon described in Sec. 15·8, an overload integrator is more suitable than a fuse. It may be a thermal circuit breaker to which heat is applied at a definite rate during the time that the overload relay has the high voltage turned off, or a slow-return synchronous timer which integrates the number of overloads per unit time in a similar way. When the number of allowable overloads per minute is exceeded, the integrator permanently disconnects the high voltage and sounds an alarm.

The preferred type of overload relay is the so-called "slow-action, slow-release" slug relay, which has a heavy short-circuited turn at the armature end of the coil. These relays can operate on about 5 ma or more, with an operating delay of a few hundredths of a second and a release or reclosing delay of as much as 0.3 sec. The slow-action relay is preferred to the rapid-action, slow-release type, because the rapid-action relay is too sensitive to small transient overloads that actually do no harm. The slow-action type, on the other hand, trips only on overloads of sufficient duration to be serious. It also ensures that the slug has sufficient circulating current to hold the relay open for its normal length of time following the overload. Because, in large beacons, it is desirable to keep the high voltage off for more than 0.3 sec following each overload, the slug relay is used to release a quick-return timer which delays the reclosing of the high voltage by about 3 sec. This synchronous-motor timer, in addition to releasing the high-voltage primary relays, applies power to the overload integrator during each 3-sec return period. If 10 overloads take place within a period of approximately 30 sec, the integrator trips and removes the high voltage until it is reset manually. A power-control system illustrating the use of such overload devices is described in Sec. 15·8.

*Cathode Preheating.*—Preheating of cathodes is necessary with a number of the higher-voltage vacuum tubes used in beacons, and especially with some of the large constant-current modulator tubes like the 715C. These tubes spark over badly unless the cathode is heated for 3 to 5 min before applying plate voltage. Gas-filled tubes, like those used in low-impedance modulators, also must be preheated properly to avoid damaging the cathode by applying the plate voltage too soon.

Small beacons that use gas-filled modulator tubes have either manual control of filament and plate voltage, or, better, automatic time delay. Time-delay relays for direct current are not entirely satisfactory. Another safety device used consists of a small relay whose coil is placed in series with the receiver plate supply in order to delay application of the modulator plate voltage until the receiver cathodes are hot. This method is not completely satisfactory, because the modulator-tube cathode usually requires a longer preheating time than the receiver-tube cathodes do, and so it is not sufficiently heated when the relay closes. One type of small d-c timer is the thermal time switch; commercial models are often unsatisfactory.

*Alarms.*—Alarm circuits for beacons, used on large ground and ship beacons for sounding an alarm whenever any fuse or relay of the power-control circuits opens, consist of a pair of normally closed contacts on the high-voltage primary relay. This relay is connected in such a way that if any fuse blows or any overload relay trips, or if any access door is opened, the high voltage is turned off and the alarm circuit is completed. The circuit is usually used to control a remote battery-operated bell.

*Power-saving Circuits.*—Beacons that must operate long hours with only an intermittent traffic load can make use of a device that disconnects all power except the receiver and the cathode heaters of the other components during periods of no interrogation. Such circuits use a relay that operates from the receiver output. The relay turns on the rest of the beacon for a period of 1 or 2 min for each single pulse received. One such circuit is shown in Fig. 15·2.

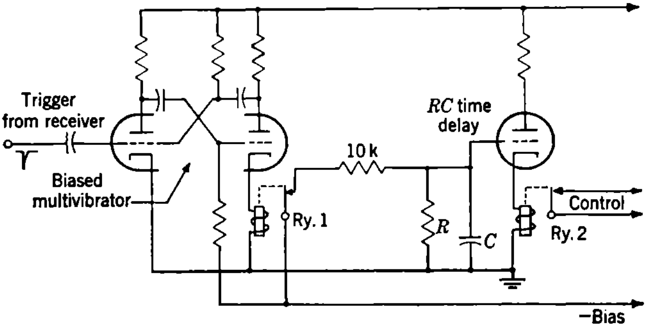


FIG. 15·2.—A circuit which disconnects the power from the plates of circuits beyond the receiver when no interrogations are being received. *R*, *C*–time-delay network; *RC* is about 10 sec.

Here, a single received pulse triggers the biased multivibrator, which, in effect, "stretches" the pulse to operate relay Ry. 1; this in turn cuts off a resistance-capacitance time-delay relay tube, which turns on the power by means of Ry. 2. If interrogations continue to arrive, the beacon remains on. If they stop coming, the RC tube will turn the power off after a minute or two.

## EXAMPLES OF POWER SUPPLIES AND CONTROLS OF TYPICAL RADAR BEACONS

The general specifications and details of the power supplies and control circuits of four representative beacons are described below.

**15·8. AN/CPN-6: A Large A-c 3-cm Beacon for Ground or Shipboard Installation.**—This microwave radar beacon is designed for use with airborne radars. A photograph is shown in Fig. 15·11, Sec. 15·12. The beacon requires 2 kva of power at 115 or 230 volts, 50 to 70 cps, single phase.

The circuit diagram of the low-voltage supply is shown in Fig. 15·3. An inspection of this diagram will show that it is a very common type of regulated supply, except, perhaps, for the feedback resistor labeled 2.2 M. This positive feedback improves the regulation with respect to fluctuating line voltage.

The medium-voltage supply is similar to the low-voltage supply, except that it is used at higher voltages and lower currents. Its circuit is shown in Fig. 15·4. The only unusual thing about this supply is the multiplicity of output voltages: eight d-c voltages are available.

The high-voltage circuit is a conventional voltage-doubler circuit. This circuit is shown in Fig. 15·5. The power supply incorporates a tapped bleeder to supply a voltmeter. Also, the 7500-ohm modulator plate-load resistor is built into the power supply proper. Note that the special condenser is built into one can with three terminals insulated and with internal balancing resistors around each section of the condenser.

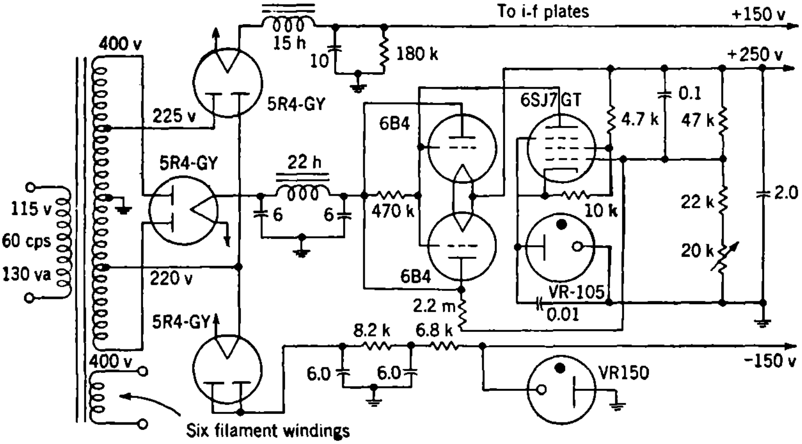


FIG. 15·3.—Low-voltage supply of a large 3-cm a-c-operated ground beacon (AN/CPN-6). Capacities are in $\mu$f.

The power-control circuit is shown in Fig. 15·6. In this circuit, when the high-voltage switch is on "automatic" and the main switch is closed, all filaments, blowers, and heaters, and all power supplies except the high-voltage supply, are turned on. After 120 sec, the timer Ry-1 supplies voltage to the 2-sec timer Ry-2. Two seconds later, Ry-2 energizes Ry-3, which turns on the high voltage. When a high-voltage overload occurs, the slug-type overload relay Ry-4 opens and disconnects Ry-2 and Ry-3 for 0.3 sec, thereby turning off the high voltage. During the 0.3 sec, Ry-2 drops back to zero. When the overload relay Ry-4

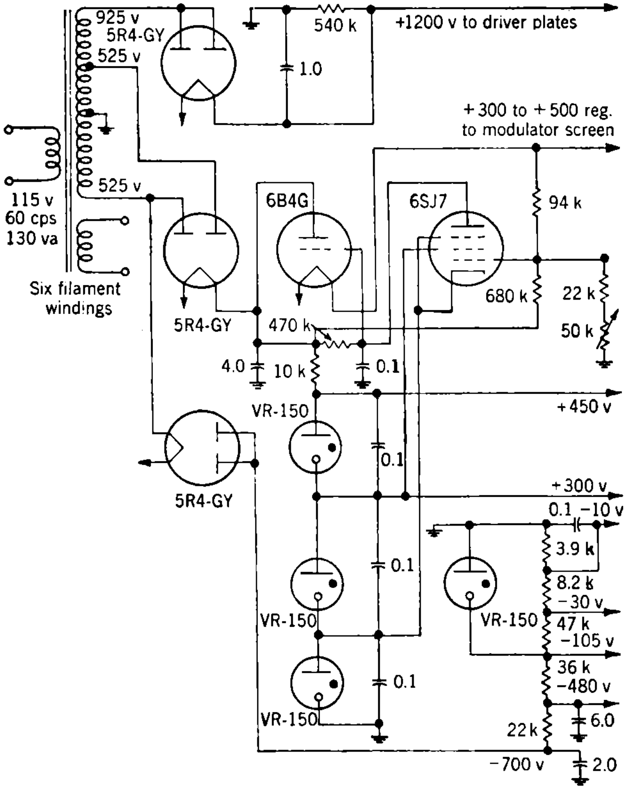


FIG. 15·4.—Medium-voltage supply of a large 3-cm a-c-operated ground beacon (AN/CPN-6). Capacities are in μf.

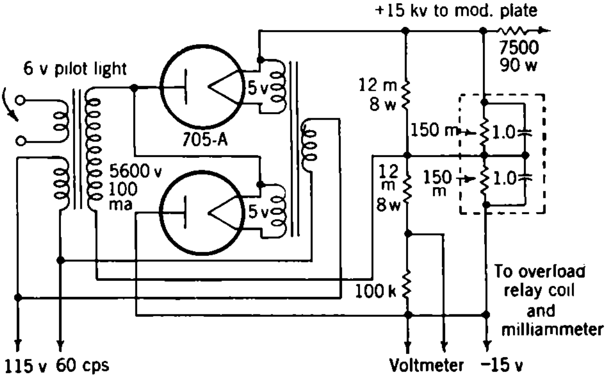


FIG. 15·5.—High-voltage (15 kv) supply for a large 3-cm a-c-operated ground beacon (AN/CPN-6). The 7500-ohm resistor is the modulator plate-load resistor.

recloses after disconnecting the high voltage, Ry. 2 starts its 2-sec cycle. During this 2-sec period, the 20-sec slow-return timer Ry. 5, the overload integrator, is energized, but advances only 10 per cent of its total travel. If the overload continues, the cycle is repeated, each time advancing the 20-sec timer Ry. 5, until after 10 to 15 overloads, Ry. 5 breaks its contacts, thereby releasing Ry. 2 and disconnecting the high-voltage power until the manual reset button is pushed.

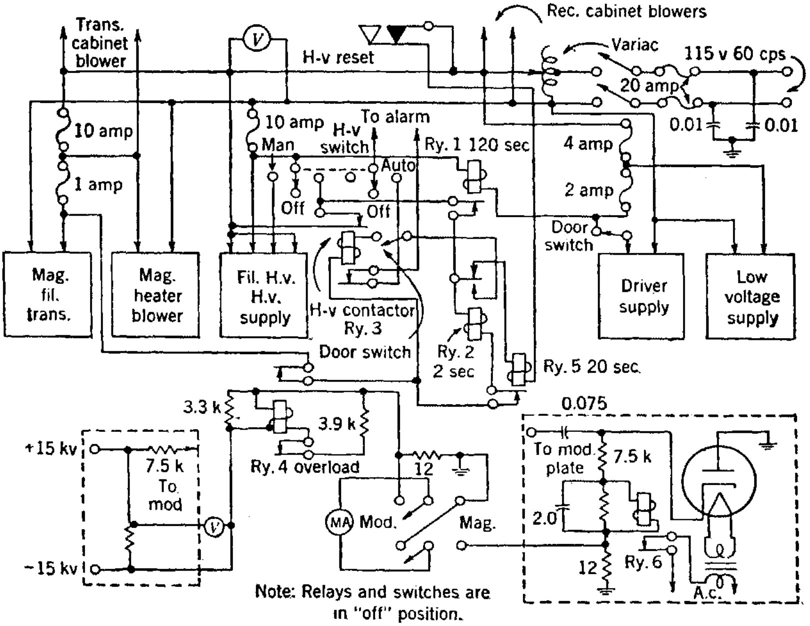


Fig. 15·6.—Power-control circuit for a large 3-cm a-c-operated ground beacon (AN/CPN-6). Ry. 1 = 120-sec. sync. timer, rapid return; Ry. 2 = 2-sec. sync. timer, rapid return; Ry. 3 = power contactor; Ry. 4 = slug-type slow-release overload relay; Ry. 5 = 20-sec. sync. timer, slow return; Ry. 6 = sensitive relay, operates 8 to 10 ma.

Note that if any of the seven fuses blows, the high voltage will be disconnected and the alarm sounded. The manual high-voltage switch is provided to make it possible to bypass all relays.

**15·9. AN/UPN-2: Lightweight A-c 10-cm Ground Beacon.**—AN/UPN-2 weighs 50 lb, exclusive of antenna and prime power source. The power required by the beacon is 150 va at 115 or 230 v ± 10 per cent, 50 to 2400 cps, single phase. Its high-voltage and low-voltage power supplies are combined. The circuit is shown in Fig. 15·7. The high-voltage supply is a conventional electronic regulating circuit, making use of the VR-tube output voltage of the low-voltage supply as a reference voltage. The overload relay is of the rapid-action slug type discussed in Sec. 15·7. The low-voltage supply is a conventional VR-tube supply with a "floating negative" to supply bias voltages. The lighthouse-tube (2C40) filament supply uses an iron-filament ballast tube. The complex resistor network is necessary to adjust the regulator and filament for optimum regulation. With this network, it is possible to adjust the voltage applied to the regulator and to shunt either the ballast tube or the lighthouse-tube filament to obtain the best possible regulation over the desired ±10 per cent voltage range.

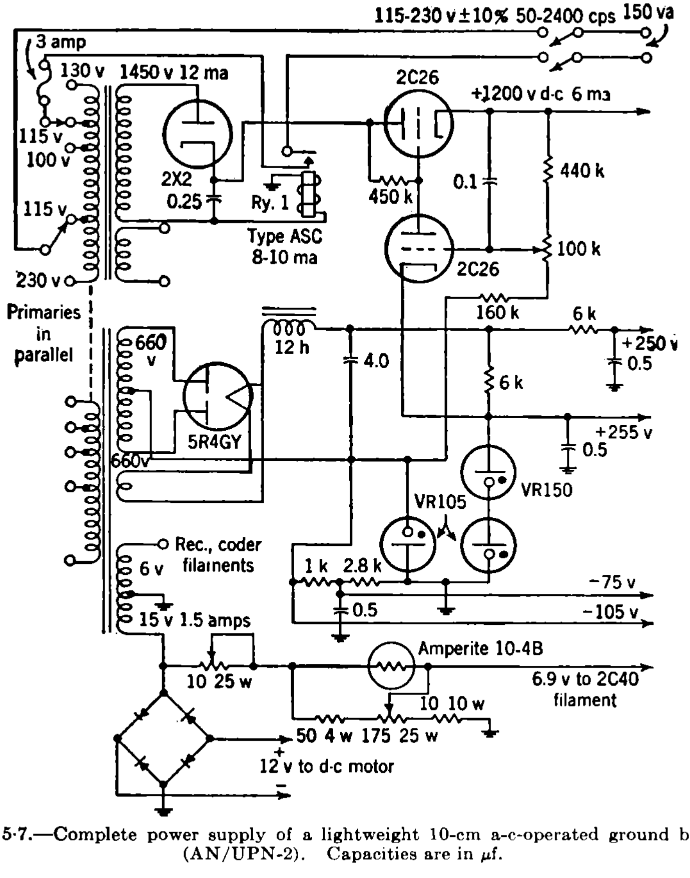


FIG. 15·7.—Complete power supply of a lightweight 10-cm a-c-operated ground beacon (AN/UPN-2). Capacities are in μf.

**15·10. AN/UPN-4: A Battery-operated "Lightweight" Beacon.**—This 3-cm beacon is probably the most complex battery-operated beacon designed and produced to date (1946). Its power supplies, therefore, are good examples of recent design. The lead-acid battery supplies 50 watts at 12 volts to the power supplies for approximately 5 hours. The power supplies are required to deliver 250 volts at 30 ma, 300 to 400 volts adjustable, 300 volts regulated, 2,300 volts at 3.5 ma, and various fixed bias voltages from 0 to −100 volts. Because the total power and number of voltages exceed the capacity of a single vibrator, two separate supplies are used in this beacon.

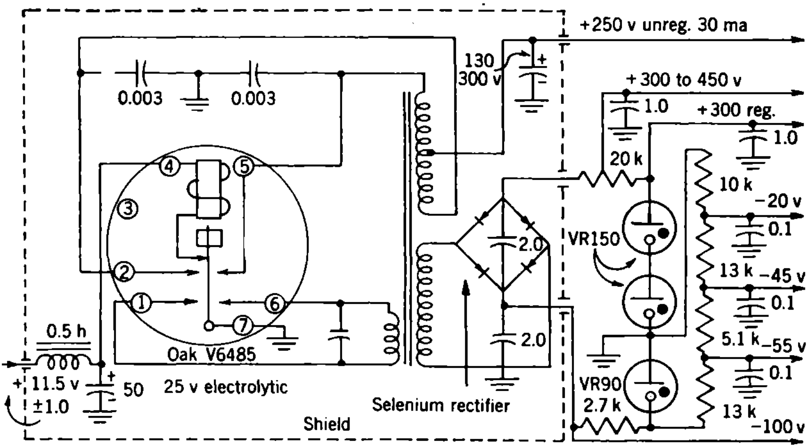


FIG. 15·8.—Low-voltage power supply for a lightweight battery-operated 3-cm beacon (AN/UPN-4). Capacities are in μf.

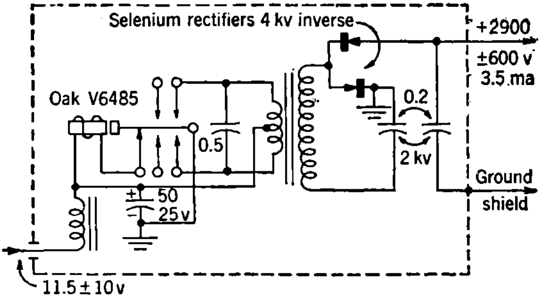


FIG. 15·9.—High-voltage power supply for a lightweight battery-operated 3-cm beacon (AN/UPN-4).

The low-voltage supply circuit is shown in Fig. 15·8. This multiple-output supply makes use of the conventional self-synchronous circuit for the unregulated 250-volt high-current output. The regulated outputs are obtained from a separate winding which supplies a selenium bridge rectifier. These rectifiers (Federal Telephone and Radio No. 23D1182), rated at 5 ma average at 65°C ambient and 1 kv peak inverse, are ½ in.

in diameter and $1\frac{7}{8}$ in. long. The bridge output supplies a resistor VR-tube voltage divider and regulator network. This power supply exclusive of VR tubes is completely enclosed in an aluminum shield can 3 in. by 7 in. by $9\frac{1}{2}$ in. The entire low-voltage supply weighs 7 lb and operates at 60 to 65 per cent efficiency.

The high-voltage supply, shown in Fig. 15·9, makes use of the same type of vibrator as the low-voltage supply but does not utilize the synchronous rectifying contacts. Instead, a voltage doubler with Federal Telephone and Radio No. 23D1101 selenium rectifiers is used. These rectifiers are also rated at 5 ma average at 65°C, but the peak inverse voltage is increased to 4 kv by increasing the number of disks. These rectifiers are $\frac{1}{2}$ in. in diameter and $4\frac{3}{4}$ in. long. The high-voltage supply is enclosed in a tight aluminum shield 4 in. by 7 in. by $5\frac{1}{2}$ in. and weighs 5 lb. The efficiency at full load is 60 per cent. Because the high-voltage supply causes less radio interference than the low-voltage unit does, it is unnecessary to use a capacitor-type bushing for the high-voltage lead.

**15·11. AN/APN-19: A D-c-operated Airborne 10-cm Beacon.**—This beacon is a small unit designed especially for installation in aircraft and use with ground radars. It operates from the 26-volt d-c aircraft supply and requires approximately 150 watts. The power supply is a carbon-pile-regulated dynamotor with a dual output. The circuit diagram of this supply is shown in Fig. 15·10.

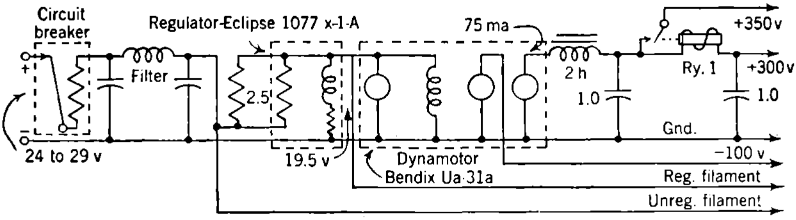


FIG. 15·10.—Power supply for a small 10-cm airborne beacon (AN/APN-19).

Although the supply, as shown in the diagram, is very simple, some difficulties were experienced with it. Among them was brush sparking, which caused radio interference. This can be ameliorated by adding a small choke and a bypass condenser to the 19.5-volt lead to the dynamotor. Also, the time-delay relay Ry-1, which closes when the receiver begins drawing plate current, does not give adequate heating time for the 2D21 gas-filled modulator tubes. In addition, operation of 6.3-volt tubes in series of three across 19.5 volts did not prove very satisfactory.

Because of these difficulties, a special dynamotor power supply was designed by Bendix. This dynamotor, with a third output of 6.3 volts for parallel heaters and with a shunt-field type of carbon-pile regulation, is a much better power supply for this type of beacon.

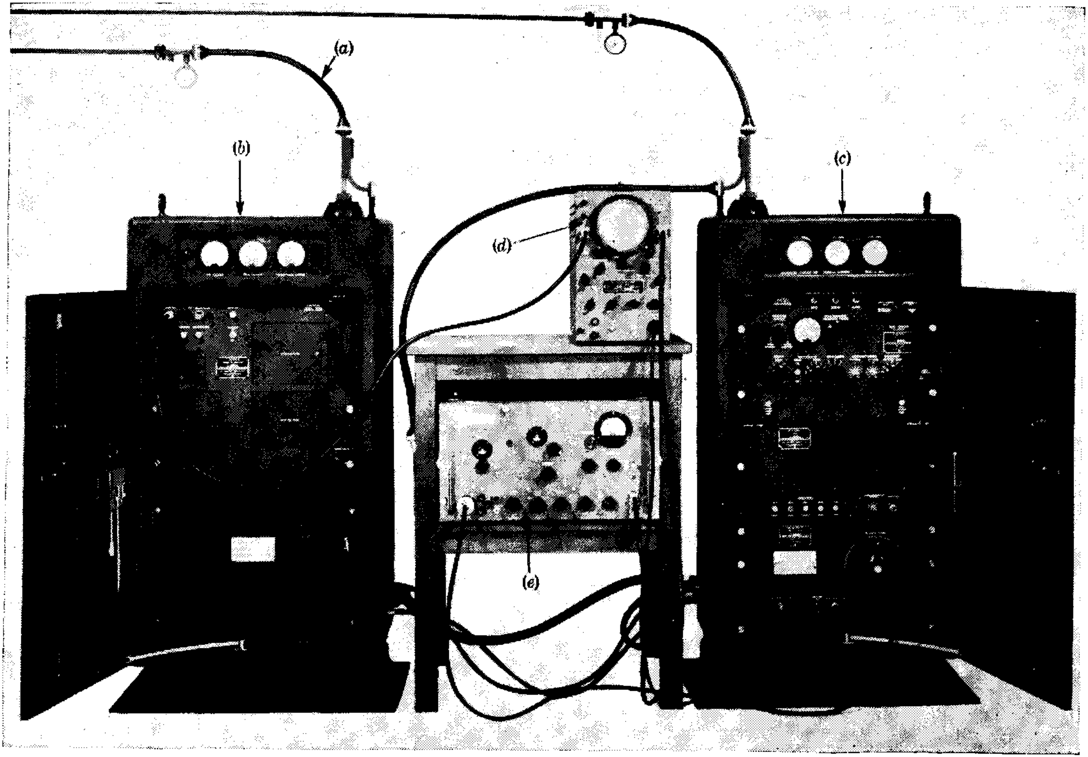


FIG. 15·11.—AN/CPN-6, a 3-cm ground beacon with external test equipment (*a*) R-f lines to antenna. (*b*) Transmitter cabinet. (*c*) Receiver cabinet (*d*) Synchroscope. (*e*) Signal generator.

## PERFORMANCE TESTING

BY M. J. COHEN AND J. J. G. MCCUE

**15·12. Monitors and Test Equipment.**—In order to adjust the various controls of a beacon and to ascertain whether its performance is satisfactory, certain auxiliary apparatus is required. If this apparatus supplies continuous information, it is called a "monitor"; if not, it is called "test equipment." A monitor is usually built into the beacon. Whether test equipment is built in or not depends largely on the allowable weight and bulk of the beacon. The two approaches are illustrated in Figs. 15·11 and 15·12. Building the test equipment into the beacon cabinet provides several important advantages: over-all compactness, convenience, and assurance that the test equipment is available. The last is the most important of the three. If a pair of beacons is installed, one to act as stand-by, some saving might result from using a single set of test equipment for both. Since, however, expensive and complex items of test equipment, like the synchroscopes and signal generators described

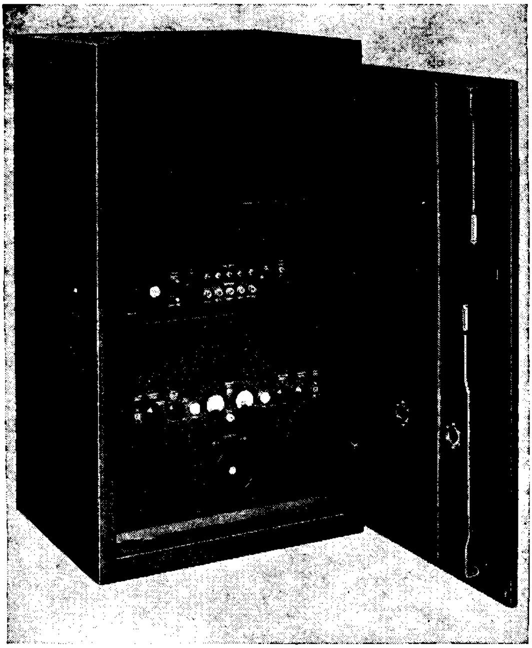


FIG. 15·12.—AN/CPN-8, a 10-cm ground beacon. The built-in test equipment, called the monitor unit, is shown in Fig. 15·13 and does not appear in this photograph. It is inserted in place of the blank panel.

below, are vital to the daily maintenance of the beacon, and are at least as liable to failure as the beacon, it is desirable to provide spares. On the whole, it is certainly advisable to build in as much of the specialized test equipment (as opposed to general-purpose test equipment like tube testers, etc.) as is consistent with over-all size and weight limitations on the individual packages.

Figure 15·13 shows the so-called "monitor" unit of the AN/CPN-8, a medium-power 10-cm ground beacon. It contains an r-f signal generator, an r-f wattmeter for setting the signal generator level and for measuring the transmitted power, a standard cavity for setting the transmitter frequency, and a synchroscope and trigger generator for viewing waveforms throughout the beacon.

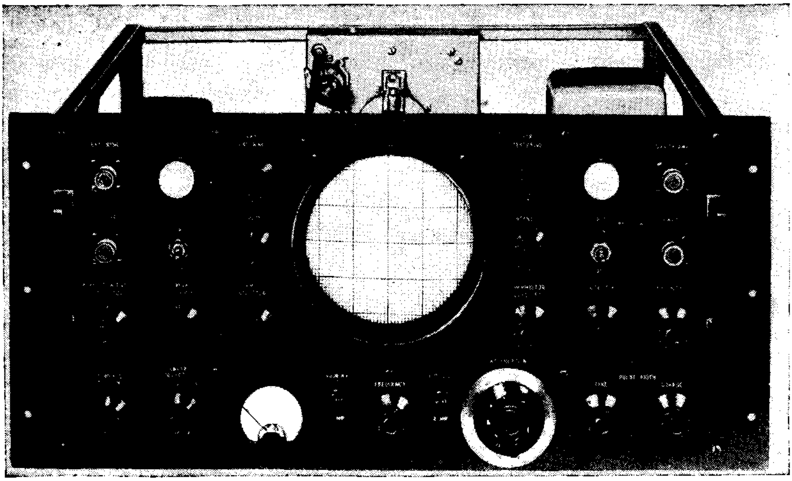


FIG. 15·13.—The "monitor" unit of the AN/CPN-8.

## TEST EQUIPMENT

The following sections describe the functions and general principles of operation of the more important pieces of test equipment[1] required for beacon operation.

**15·13. The Signal Generator or "Test Set."**—The usual signal generator for testing beacons contains—

1. A low-power oscillator tube, tunable over the frequency range of the receiving and transmitting bands of the beacon.

---

[1] The actual design of test equipment is discussed in detail in *Technique of Microwave Measurements*, Vol. 11, Radiation Laboratory Series.

2. A circuit for pulsing the oscillator tube, driven either by an externally supplied trigger or by an internal trigger source.
3. Provision for coding the output pulse in accordance with the interrogation codes used in the system, so that the operation of the beacon decoder can be tested.
4. A device, such as a thermistor bridge, for absolute measurement of r-f power at low levels, and a power-adjustment control to enable the pulse power output to be adjusted to a standard level.
5. A wide-range calibrated attenuator for adjusting the pulse power output over a range suitable for testing the beacon receiver.
6. A frequency meter for use in setting the frequency of the output.

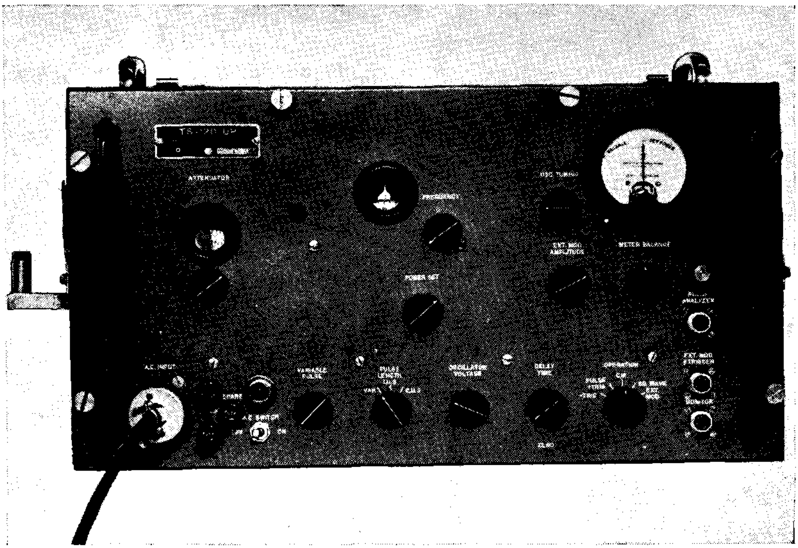


FIG. 15·14.—The TS-120/UP, a 3-cm test set.

Provision is frequently made for measuring the power level of an external source, such as a beacon transmitter, by use of the same power-measuring device. The signal generator is usually called a "test set." A 3-cm test set, the TS-120/UP, is shown in Fig. 15·14.

**15·14. The Synchroscope and Envelope Viewer.**—The synchroscope is a cathode-ray oscilloscope with a horizontal sweep initiated by a trigger circuit. The phase of the sweep with respect to the trigger is adjustable. The sweep speed is also adjustable; in current designs it usually has an upper limit of about 1 in./$\mu$sec. The sweep speed can be measured by a built-in timing circuit that provides accurately timed calibrating marks. For vertical deflections, the signal is usually applied directly to the deflecting plates. Many synchroscopes contain a crystal detector that can be coupled to the vertical-deflection plates through a moderate-gain broadband video amplifier. This accessory, called an "envelope viewer," is useful for observing the shape of pulse envelopes. Figure 15·15 shows the TS-143/CPM-1, a synchroscope designed for use with beacons.

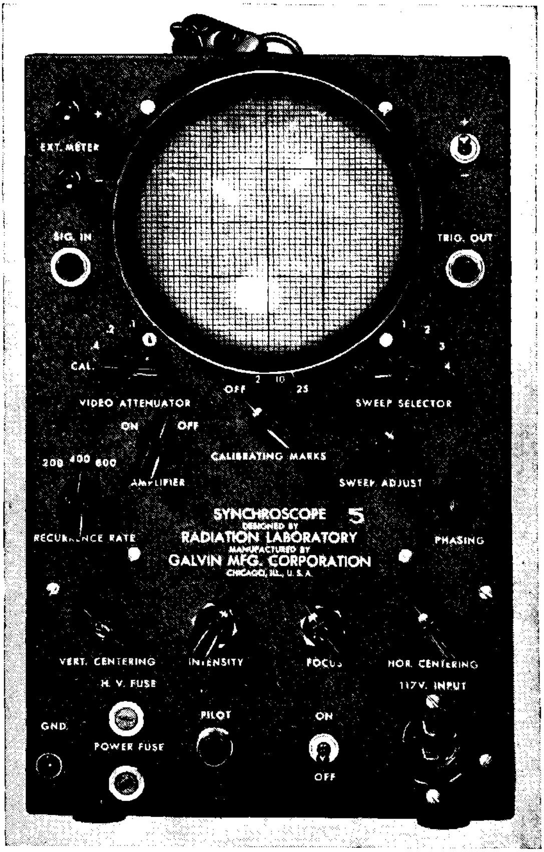


FIG. 15·15.—The TS-143/CPM-1, a synchroscope designed for beacon use.

**15·15. The Spectrum Analyzer.**—The high-frequency spectrum analyzer is essentially a superheterodyne receiver with a frequency-swept local oscillator and a narrow-band i-f amplifier. The output of the receiver produces a vertical deflection on an oscilloscope, while the horizontal sweep on the oscilloscope is the same sawtooth wave that sweeps the frequency of the local oscillator. The pattern on the oscilloscope screen is therefore a graph of output voltage as a function of fre-

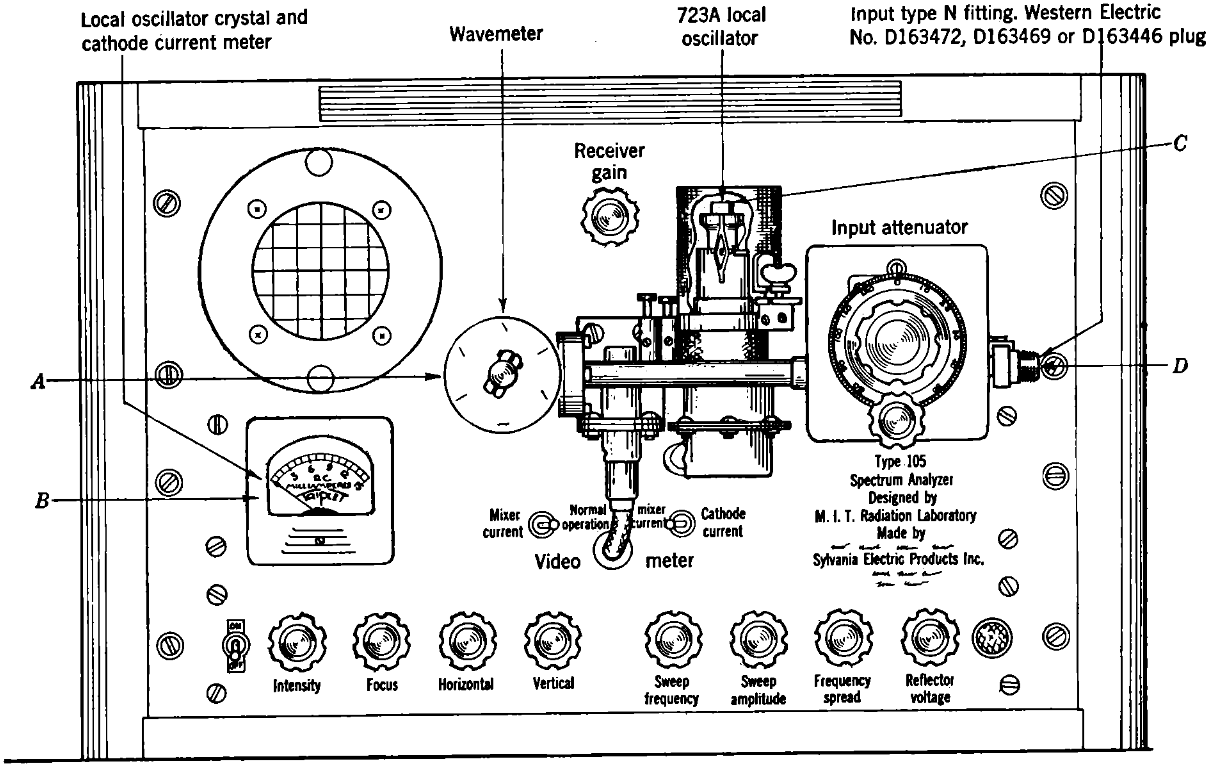


FIG. 15·16.—A 3-cm spectrum analyzer (type 105).

quency. The distribution of power in a pulse of any given shape can be calculated by Fourier analysis. By examining the r-f pulse with a spectrum analyzer, it can be determined whether the actual shape approximates the desired shape, and, in particular, whether the r-f pulse shape is approximately rectangular, whether its width is correct, and whether the frequency of the pulse is centered at the proper value. A spectrum analyzer is a convenience for beacon operation but not a necessity; the necessary functions it performs can be carried out by other means. Figure 15·16 shows a 3-cm spectrum analyzer.

**15·16. Wavemeters and Reference Cavities.**—Ultrahigh frequencies can be measured by means of a heterodyne wavemeter, which consists of a variable-frequency oscillator, a detector amplifier, and a pair of headphones. The detector amplifier permits the operator to adjust to zero beat between the oscillator and the signal by matching the oscillator frequency to the signal frequency. The oscillator is carefully calibrated, and the calibration can be checked by beating the oscillator against a harmonic of a low-frequency (5-kc) crystal-controlled oscillator, which is built into the equipment. The high-frequency oscillator can also be used as a signal generator for other tests. This technique can be used at microwave frequencies when very high precision is required.

At microwave frequencies, the coaxial wavemeter is less accurate but very useful. In the reaction-type coaxial wavemeter, a small amount of power is coupled through a probe into a section of coaxial line, short-circuited by a plunger. By means of a precision screw, the plunger can be moved along the line until resonance causes a change in the output voltage of a crystal rectifier connected to the probe. The distance between two successive resonating positions of the plunger is half a wavelength. In transmission-type coaxial wavemeters, the crystal detector is connected to a separate probe, and its output voltage is zero except when the plunger is at or close to one of the resonating positions. An advantage of the reaction type is that its output voltage is different from zero whenever the oscillator is functioning, but the transmission type exhibits a greater change in output voltage when the plunger passes through a position of resonance. Fluctuations in oscillator output are therefore less likely to mislead the operator if a transmission-type wavemeter is used.

Coaxial wavemeters are most useful in the wavelength range from about 5 to about 50 cm; rough measurements at shorter wavelengths are possible.

In the 1-cm to 10-cm region, resonant cavities with adjustable tuning plungers are used as wavemeters in conjunction with crystal detectors. Both reaction and transmission types are available. Their operation is similar to that of coaxial wavemeters, but they are superior in applications requiring temperature compensation, high selectivity, or high precision, and are more convenient to use because only one reading is required. Cavities with dials reading directly in frequency or wavelength are available.

No tunable wavemeter has yet been able to meet the requirement of fixing the beacon frequency in the 3-cm range to about ± 1 part in 10,000. Transmitters in beacons operating in this range are, therefore, monitored by fixed-tuned resonant cavities with $Q$'s of about 2000, which have negligible temperature coefficients and are hermetically sealed to prevent the resonant frequency from changing with humidity. For protection and convenience, such reference cavities are usually built into the beacon. A fixed probe or a directional coupler takes a small fraction of the transmitter power, which passes through the standard cavity and is detected by a crystal. The output of the crystal is sometimes connected to a meter through a d-c amplifier. More commonly, however, the connection is through an audio amplifier which covers the range of the pulse-repetition frequencies.

The local oscillator in microwave beacon receivers is usually set to the proper frequency with the aid of a standard cavity; when possible, this should be accomplished by AFC. If AFC is not feasible, the standard cavity is used in the same way as the transmitter reference cavity. In this case the amplifier may be unnecessary, because the local oscillator is not pulsed and the average input power to the crystal is large enough to give a reading on a milliammeter connected directly to the crystal. The local-oscillator frequency need not be adjusted with the precision required in the case of the transmitter frequency, because the i-f band coverage usually can be made enough broader than the required coverage to allow for some inaccuracy in centering.

**15·17. Measurement of R-f Power.**—At microwave frequencies, r-f power is almost always measured by means of its heating effect. The most direct method for high power levels is to measure the temperature rise of a stream of water in which the power is absorbed; this method is widely used in the laboratory but seldom in the field.

The most common devices for use outside of the laboratory are based on the change of resistance of a suitable material with temperature. A thermistor, or other bolometer, in one arm of a Wheatstone bridge can be matched into a waveguide or other transmission line so that it absorbs practically all of the power in the line. The r-f power is measured by comparing it with the d-c power required to maintain the thermistor at the same temperature. Present thermistors operate in the range from 10 μw to 10 mw. Higher powers are measured by interposing a calibrated attenuator between the oscillator and the thermistor.

It is possible to measure the power in a waveguide without absorbing much of it by using a Johnson meter, which consists essentially of a short section of waveguide made of thin constantan. Radio-frequency power transmitted through the guide is, in part, dissipated in the poorly conducting constantan, thereby heating it. The temperature rise in the constantan is measured by means of a resistance thermometer and is proportional to the power transmitted through the waveguide. Average powers of a few tenths of a watt of 3-cm radiation can be measured; the meter is less sensitive at longer wavelengths. Since the special section of waveguide absorbs only a small fraction of the power in the guide, it can be installed as a permanent part of the r-f line. The Johnson meter has the further advantage of being rather insensitive to frequency change because no matching is necessary, but this advantage is not likely to be significant in beacon applications.

## MONITORS AND BUILT-IN TEST FACILITIES

**15·18. Test Jacks.**—The beacon should have built-in jacks at several points. Such jacks facilitate the measurement of voltage and the viewing of waveforms with a synchroscope, simplify routine checking, and permit quick location of trouble in the beacon. It is useful to provide jacks for testing the outputs of the receiver, discriminator, coder, and modulator, and the transmitter current. The circuits should include voltage dividers where necessary, so that the synchroscope can be applied, for example, directly to the test point for the modulator output.

In addition to the test points listed above, r-f test points must be provided in both the receiver and the transmitter lines. The former are used for measuring receiver sensitivity and the latter for measuring transmitter power. Since these test points should incorporate fixed and known attenuation, directional couplers are well suited for the purpose. They need not be built into a beacon designed for maximum portability provided the antennas are readily removable so that the test equipment can be applied directly to the antenna connections. The receiver-input test point can be used to introduce into the receiver, by means of a signal generator, an r-f signal that is initiated by the synchroscope trigger and is therefore in the proper phase with the synchroscope sweep. The synchroscope is then applied to the various test points.

It is convenient to locate all the test jacks in one accessible location, but care must be taken to prevent undesirable interaction of the leads. The lead from the receiver output must not place enough capacitance across the output terminals to distort the pulse, and must not feed any of the output voltage back into the input terminals. Isolating this lead by means of a cathode follower is sometimes desirable.

The viewing of the current pulse in the magnetron is accomplished by placing a suitable resistor (10 to 100 ohms) between the anode and ground. The test lead runs to the magnetron anode through a suitable matching resistor. The trouble of insulating the anode from ground will usually be well compensated for by the ease and assurance with which both magnitude and duration of the pulse of current in the magnetron can be adjusted and observed. Any malfunction of the magnetron resulting from aging, mismatch in the antenna system, or other cause, is indicated unmistakably by the appearance of the current pulse.

A resistor placed in the cathode lead of a hard switch-tube will produce a pulse of voltage which can be correlated rather directly with the magnetron pulse. This method can be used when it is essential to ground the anode of the magnetron.

**15·19. Audio Monitors.**—Because pulse-repetition frequencies used in interrogators lie in the audible range, it is possible, with the aid of audio amplifiers, to monitor the operation of the beacon aurally. The information may be made available at the beacon and at remote monitoring points. Monitoring the receiver or coder output enables one to determine whether pulses are reaching the beacon and whether they are interrogating it. A crystal detector used with a resonant circuit permits aural monitoring of the transmitter frequency.

When the coder is operating, the blanking gate provides a signal of relatively high power level, because both the voltage and the duration of the gating pulses are considerable. A cathode follower driven by the blanking gate will often give sufficient power to operate a headset. If there is no coder, the blanking gate for the receiver can be used, but some amplification before the cathode follower may then be necessary.

The signal from the crystal detector of a transmitter monitoring cavity is necessarily weak, because the crystal will burn out unless there is considerable attenuation between it and the transmitter. This signal must, therefore, be amplified perhaps as much as 30 db. It is desirable to approach a match of the crystal detector to the grid of the first amplifier tube by using a stepup transformer of high ratio. A microphone input transformer with a stepup ratio of 1:20 to 1:50 is often satisfactory. Because the resistance of the crystal is lowered by the r-f signal, a condenser in parallel with the crystal charges rapidly with the pulse and discharges slowly after the pulse. This widens the voltage pulse on the grid of the first tube, and therefore increases the average power available in the amplifier output. The input circuit must be carefully shielded to minimize pickup of video signals from the modulator or elsewhere, and to prevent r-f power from entering the crystal detector without passing through the resonant cavity. The frequency characteristic of the amplifier does not have to be very flat, because only the fundamental—and perhaps the first few harmonics—of the interrogating pulse-repetition frequency need be amplified. If the monitor is to be switched successively to different test points in the beacon, padding should be introduced to bring all of the output signals to approximately the same level when the beacon is functioning properly, to prevent inconvenient variations in the output level when the monitor is switched from one test point to another. A milliammeter in the output circuit of the amplifier is sometimes useful to provide a visual indication.

There should be sufficient decoupling in the input circuit so that the amplifier will not affect the circuits that it is monitoring. The power-supply leads to the amplifier may require decoupling. Careful grounding, perhaps through a common ground, may be needed to avoid pickup and regeneration arising from currents circulating in the chassis.

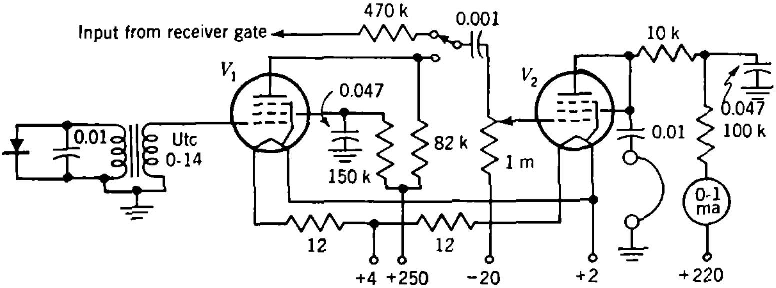


FIG. 15·17.—An audio monitor for a portable beacon.

Figure 15·17 shows the circuit of an audio monitor for a portable beacon.

**15·20. Other Test Facilities.** *Built-in Meters.*—It is desirable to show the power-line voltage, the high voltage for the modulator, and the transmitter current, on built-in panel meters. As mentioned above, it will often be convenient to read the output signal of the audio monitor on a meter.

*Other Test Equipment.*—Items of test equipment that are not built in but should be available for maintenance work, include the following: a well-matched dummy load which can dissipate the maximum transmitter power, for use when the antenna mismatch is suspected or for tests on the beacon transmitter when it is not desirable to radiate high power; a capacity-voltage divider, if one is not built in, for viewing the transmitter voltage pulse; and standard items such as a volt ohmammeter, tube tester, and a fluxmeter (if the beacon uses a magnetron).

# CHAPTER 16

# BEACON SYSTEM SYNTHESIS

By B. W. Pike, J. J. G. McCue, and M. J. Cohen

Beacon synthesis here is taken to mean the part of beacon designing that remains after over-all performance specifications have been determined by the methods outlined in Part I and the required major components have been designed as described in the previous chapters of Part II. The remaining problems of design are control of the over-all delay of the beacon, interaction of the components, prevention of radio noise, suppression to prevent triggering by near-by radar sets, protection of the crystal, and packaging.

## 16·1. Beacon Synthesis.[1]

—Because beacon synthesis is something of an art, it is difficult to analyze the methods used by designers. Usually, after specifications of the entire beacon-interrogator system have been decided upon, the characteristics desired for the interrogator and for the beacon are outlined in detail. The beacon designer then starts with consideration of the specifications for over-all performance, weight, size, and shape, which are determined by the specific tasks the beacon is to perform. From these specifications, the specifications for the performance of the major components are calculated, and a guess made as to the proportion of the available space and weight to allow to each major component. At the same time, a rough idea of the desired shape of each major component may be formed, in view of the desired final shape of the beacon.

The second step is to reduce the major component specifications to practice in the form of "breadboard" circuits or models. When all major components are operating individually, the critical third step is made.

The third step, assembly of the beacon, is the test of the adequacy of the major component specifications, for it is here that over-all delay, interaction of components, and radio noise problems first appear. It is a rare designer indeed whose beacon immediately functions properly at this step.

After the components have been made to work together properly by modification of the circuits, shielding, filtering, and so on, the fourth step

---

[1] Sections 16·1 to 16·9 by B. W. Pike.

is to package the components into an integrated beacon. After a packaged prototype is built, new problems of component interaction, noise, packaging, and the like, usually arise. The final step is the correction of these unforeseen difficulties.

After all these steps have been taken, it is usually found that the beacon meets performance specifications, but is just a little larger and heavier than originally planned.

**16·2. Interaction of Components.**—The subject of component interaction, exclusive of radio noise, will be discussed by considering each major interconnection of components and pointing out the known interactions and their cures. Cross interactions will be considered later.

*Receiving Antenna to Receiver.*—Difficulty is experienced here when a tuned receiver input is affected by the reactance of a mismatched antenna. This effect is aggravated when the antenna-to-receiver line is many wavelengths long, because small temperature changes can then cause changes in the line length which may result in considerable detuning of the receiver. The obvious cure is a well-matched antenna and line. If necessary, a phase shifter can be adjusted manually to present the desired impedance at the receiver.

*Receiver to Decoder.*—Few troubles are met here if the output circuits of the receiver were designed with the decoder in mind. The designer should be sure that a signal of adequate amplitude and shape from a source of the proper internal impedance is delivered to the decoder. Care must be taken to prevent reaction on the receiver from the decoder when signals are rejected. Decoder-video reaction from accepted signals is not important. Because decoders are often quite sensitive to supply voltages, the receiver should not affect the power supply of the decoder in any way.

*Decoder to Blanking Gate and Coder.*—The possibility of interaction among these components is small, because the blanking gate and coder go into action only after the decoder has completed its work. The end of the blanking-gate pulse, however, must not be allowed to react on the decoder.

*Coder and Blanking Gate to Modulator.*—Here again it is important to match the coder output to the required modulator input, remembering that the modulator must be capable of responding to the closely spaced pulses of the code. Although coder circuits are relatively insensitive, there is a possibility that video pulses from the modulator or transmitter r-f pulses at the beginning of the code can interfere with the formation of the remainder of the code. Also, the trailing edge of the blanking-gate pulse must not be allowed to reenergize either the coder or the modulator.

*Modulator to Transmitter.*—The major precaution required here, assuming proper matching of the modulator output to the transmitter tube, is the necessity for providing for the very large transient voltages or currents produced by spark-over of either the modulator tube or the transmitter tube. If the transmitter is not matched to the modulator, reflection of the modulator pulse may occasionally produce a second unwanted transmitter pulse.

*Transmitter to Antenna.*—Here a problem similar to that encountered in the antenna-to-receiver link exists. It is usually more serious in transmitters, because a mismatch can pull the oscillator frequency enough to cause serious trouble. As in the antenna-to-receiver link, the designer can match the load and line or use a phase shifter. With transmitters, control of the frequency by an AFC system or a stabilizing tuner helps to minimize this difficulty.

*Cross-interactions: Transmitter Antenna to Receiver Antenna.*—This linkage can cause two types of trouble. The first occurs when enough power is coupled into the receiver from the transmitter antenna to damage crystal detectors. This problem is solved by proper design and location of antennas or by duplexing when a single antenna is used. The second difficulty arises when the strong transmitted signal blocks the receiver, or when reflected transmitter signals return to the receiver after a delay long enough to allow them to reinterrogate the beacon. The reinterrogation problem is solved by the methods described in Sec. 6·9. The problem of receiver blocking must be taken care of in the design of the receiver; when a blanking gate is used, the receiver is allowed some time to recover.

*Miscellaneous Cross Interactions.*—Common coupling through power supplies can cause difficulties. For example, in an airborne beacon whose transmitter and receiver are supplied by a dynamotor, the receiver sensitivity may exhibit an oscillatory variation caused by a drop in its plate voltage when the modulator is triggered. Elimination of any common power-supply impedance, such as a common filter choke, usually solves this problem.

Some receivers and transmitters are microphonic. Care must therefore be taken to see that blower motors and the like do not cause interfering vibrations.

From the above, it can be seen that troubles with component interaction are likely to arise from problems which are obvious in retrospect but are sometimes overlooked in design.

## OVER-ALL CONTROL OF THE DELAY

**16·3. General Considerations.**—Beacon delay may be defined for convenience as the time lag between the reception of an r-f interrogating pulse and the transmission of the first r-f reply pulse by the beacon antennas. In the case of coded interrogations, the r-f interrogating pulse is taken to be the one which elicits the response; it is usually the one which triggers the range circuits of the interrogator. Beacon delay exists because of the finite time required by circuits used in beacons to respond to pulses impressed upon them and because of time for propagation in transmission lines.

Good beacon design, in general, keeps the absolute over-all delay from r-f reception to r-f transmission at a minimum. In addition, variation or change in delay with such factors as signal strength of the interrogating pulse, or line voltage fluctuations, should be kept at a minimum. Pulse-to-pulse changes in delay, which are evidenced as jitter in the reply pulse of the beacon, should also be minimized.

The exact amounts of over-all delay, variation in delay and jitter that can be tolerated will depend largely on the application of the system, and the precision of the range information required. In many cases, it is more desirable to have a considerable over-all delay that is known and constant than to have a comparatively short delay which exhibits such variation or jitter as to make its precise value uncertain. In normal beacon navigation, a delay of about 5 μsec is of small consequence, even though it constitutes an error of ½ mile in range. Not only is this small in comparison to any but the shortest sweep length on a typical airborne radar, but it can be allowed for readily. Here, too, as much as 0.25-μsec jitter goes unnoticed by the radar operator.

**16·4. Magnitude of Delay.**—An estimate of the magnitude of the time delays to be expected from various circuits used in beacons is given below. This should make clear the nature of the problem presented and indicate where the greatest delays and delay variations are to be expected.

*Receivers.*

1. *Wideband Video* ($\approx$ 1.5 Mc/sec). The delay introduced is small for all signal strengths (0.1 μsec). This applies to all receiver types.
2. *Narrowband Video.* For very weak signals, the delay introduced here may equal the width of the interrogating pulse width. The delay occurs because the beginning of the receiver output pulse is not steep. This results in a variation with signal strength in the time required to reach some threshold value. When signal level is about 6 db or more above the minimum for triggering the delay decreases considerably, to about 0.5 μsec.
3. *Superregenerative Receivers.* Superregenerative receivers are subject to a variable delay which takes any value between zero and the quench period. (See Sec. 8.6.) This delay is different in a random way from pulse to pulse, and thus constitutes a jitter as well.

4. *Low Intermediate Frequency.* In some beacon receivers, very low intermediate frequencies, down to 1 or 2 Mc/sec, are used. The effect of this is similar to that produced by the quench frequency in a superregenerative receiver. Both delay and jitter of an amount that is an appreciable fraction of one period in the intermediate frequency are introduced.

*Decoders.*—The delay depends upon the type, as follows.

1. *Pulse-width discriminator to reject pulse over* 2 *μsec long.* This necessarily produces a delay of about 2 μsec, sometimes slightly more.
2. *Pulse-width discriminator to reject pulse over* 5 *μsec long.* The delay is equal to the pulse width for some designs. It is maintained at a constant value of 5 μsec for other designs.
3. *Coincidence.* The delay depends on the circuit but is generally quite small (less than 0.25 μsec).
4. *Double-pulse.* The delay is usually less than 0.25 μsec with respect to the second pulse.

*Duty-ratio Limiting Gate (Blanking Gate).*—This circuit normally produces a delay less than 0.2 μsec. The circuit can cause serious jitter, if poorly designed, particularly near the maximum duty ratio.

*Blocking Oscillators and Multivibrators.*—The delay is small here, being about 0.1 to 0.2 μsec, including the regeneration and pulse rise time, if the circuits are properly designed. If, however, even a well-designed blocking oscillator is fired by a small or insufficiently sharp trigger, the regeneration or starting time may increase to 0.5 to 1.0 μsec, although the time of rise from 10 to 90 per cent of full voltage of the output waveform may still be under 0.2 μsec. In general, multivibrators produce somewhat less delay than blocking oscillators.

*Modulators.*—The delay depends on the type, as follows.

1. *Hard-tube.* This type normally produces a small delay.
2. *Gas-filled Tube.* This type produces a considerable delay, from 0.5 to 1.5 μsec, depending on the amplitude of the trigger used for the gas-filled tube and on the type of tube.

*Transmitter.*—The delay depends on the type as follows.

1. *Magnetron.* Magnetrons generally have a negligible inherent starting time (0.1 μsec or less). Frequently, however, the modulator pulse is purposely sloped to help the magnetron start oscillating in the proper mode. The over-all delay of the applied pulse plus the magnetron starting time is then about 0.1 to 0.25 μsec.

2. *Triode Oscillators.* The delay caused by the build-up time is about 0.1 to 0.3 μsec. Use of an applied pulse which is too small may increase this figure.

*R-f lines.*—The delays in r-f lines are usually small enough to be negligible except for high-precision ranging systems. The delay is readily calculated from the length of line and the group velocity of propagation in it.

Despite the manifold causes of beacon delay, it is quite possible to keep over-all delays to as little as 1 μsec by suitable design. More important, the delay jitter can be kept less than 0.1 μsec for accurate ranging. This requires wideband video circuits, and is accordingly expensive in power and weight.

**16·5. Intentional Delays.**—In some beacons a known and constant delay of comparatively great magnitude is intentionally introduced for various reasons, as follows.

1. Airborne beacons frequently must have suppressors by which they are prevented from replying to a radar in the same aircraft. A fixed delay is introduced before the circuit to be suppressed in order to give the suppressor time to work.
2. Coders and decoders generally involve intentional delay circuits, such as delay lines, gates, sloping triggers, or other delay devices (see Chaps. 9 and 10).
3. Over-all beacon delay, radio frequency to radio frequency, may purposely be made any desired value greater than the minimum inherent delay, by adding delay by various devices described below.

Added delay may be required to bring the over-all beacon delay to some fixed predetermined figure, which is monitored and accurately maintained. Fixed precise delays are needed in systems in which ranges to the beacon are measured with great accuracy for precision navigation.

Herewith are enumerated the various methods of achieving the intentional delays referred to above.

1. *Electronic delay circuits,* such as multivibrators, phantastrons, and so on, may be used to induce an intentional delay. The principle here is to use a pulse to initiate a delay circuit which produces an output signal after proper delay. The output signal is then used to form a sharp pulse by differentiation. An obvious example is the production of the second, third, and following code pips in a beacon. The same methods can be used to delay the first pip if desired. These methods are best suited for delays greater than 5 μsec.

2. *Delay lines* in various arrangements are frequently used for delays of a few microseconds. Whereas the first method mentioned above offers a continuously variable adjustment by the use of a gate whose length is continuously variable, this method offers adjustment of delay in steps by use of a tapped delay line. It is best suited for delays less than 10 μsec.
3. *Supersonic devices* can be used for very long delays. Suitably designed supersonic delay lines provide delays of the order of milliseconds, with a precision of fractions of a microsecond.
4. *Sawtooth voltages* also offer continuous control of delay. Here the bias on a stage triggered by a sawtooth waveform is controlled and varied, thus varying the time it takes to trigger the stage.

**16·6. Summary.**—The causes of delay and variation of delay are tabulated below. Causes of jitter are listed separately, although in most cases they are distinctly related to delay itself.

1. *Causes of delay and delay variation:*
    - *a.* Rise time in narrowband circuits.
    - *b.* Propagation time in r-f and video lines.
    - *c.* Time required for decoding.
    - *d.* Sloping waveform reaching some threshold point.
    - *e.* Regeneration time in regenerative stages.
    - *f.* Build-up time in oscillators.
    - *g.* Rise time in very-low-frequency i-f amplifiers.
    - *h.* Intentional delays, using delay lines or other devices.
2. *Causes of Jitter:*
    - *a.* Uncertainty of threshold levels.
    - *b.* Uncertainty in regeneration times.
    - *c.* Uncertainty in build-up time in oscillator.
    - *d.* Use of low-frequency i-f amplifiers.
    - *e.* Use of superregenerative receivers.

Both delay and jitter are cumulative from stage to stage, and their "cure" is obvious from the causes enumerated above. They can be eliminated by the use of fast-rise-time waveforms, strong triggers for regenerative stages, and the use of correct voltages and loads for oscillators. As a general rule, circuits with more inherent delay have more inherent jitter and more inherent variation of delay due to supply voltage changes and other parameter variations.

## PREVENTION OF RADIO NOISE

**16·7. Noise Triggering and Noise Radiation.**—The designer of a radar beacon is concerned with two aspects of the radio noise problem.

The first is the protection of the beacon from triggering by radio interference, coming from outside the beacon or generated by some component within the beacon itself. Second is the prevention of noise generated by the beacon from being of sufficient strength to interfere with adjacent electronic equipment.

The first problem is merely a part of making a beacon that operates properly, but the second has often been neglected, with unfortunate results. The armed services have published general specifications[1] which may serve as guides to the allowable amount of noise that may be radiated or conducted into or away from the equipment. The major points of these specifications are summarized below:

1. Conducted noise, as measured across any two terminals, including ground, shall not exceed 50 $\mu$v from 0.150 to 20 Mc/sec.
2. Radiated noise shall not exceed 5 $\mu$v from 0.150 to 150 Mc/sec when measured with an approved noise meter at 3 ft from the equipment.
3. Any device intended to be nonradiating, such as a receiver, shall not feed more than 400 $\mu\mu$w of r-f energy into its antenna. (This requirement is unnecessarily severe, and is often impossible to meet.)
4. The conducted noise susceptibility ratio (antenna input/any other input) required of any receiver shall be 1000 to 1 over the entire receiver frequency band.
5. Radiated noise susceptibility shall be such that no interference shall result from a field of 1 volt per meter of any frequency or type of modulation likely to be encountered as a result of operation near other electronic equipment.

**16·8. Elimination of Noise Produced by Beacons.**—Pulsed radar beacons are inherently prolific sources of radio interference because of the high-power video pulses generated by their modulators. The corresponding radiated energy must be confined to the beacon enclosure. The noise from other sources in beacons, such as d-c motors, vibrators, and so on, not only must be confined to the beacon cabinet, but also must be kept out of the beacon receiver.

Obviously, the receiver should be shielded and filtered to prevent noise from reaching it. The remaining components, including all noise sources, should be placed in other shielded and filtered enclosures to prevent the escape of noise energy. This procedure, although it is simple in principle, is not always easy to follow, as it is difficult to filter connect-

---

[1] U. S. Navy Dep't Specification, RE 13A 554 E, Aug. 25, 1944.

U. S. Army Air Force Specification No. 71-854, *Gen'l Spec. for Aircraft Electronic Equipment*, Sheet No. 30.

ing leads adequately over the required frequency range without introducing numerous large and complex filters. Also, in some cases, shielding of the required excellence interferes with proper dissipation of heat.

The preferred method is to concentrate on each individual generator of noise in order to decouple it from other circuits by careful placing of components, conductors, grounds, and shields. When *RC* or *LC* filtering is used with each noise generator, the noise level within the beacon cabinet should be low enough to allow the use of ordinary shielding and filtering of the receiver. When the level of internal ambient noise is low, ordinary shielding and filtering of the beacon as a whole will suffice to prevent excessive radiation. The following points will serve as a guide to these preferred methods of eliminating noise.

*Circuit Design.*—Waveforms with changes of slope that are more abrupt than necessary for proper performance should not be generated because they involve radiation of unwanted high frequencies. In general, video waveforms that have components of appreciable strength at frequencies above 10 Mc/sec are not needed in beacons and should be avoided. The usual modulator with pulses from 0.5 to 1 μsec long generates most noise energy between 0.2 and 4.0 Mc/sec.

Video pulses should be generated near the point of application and confined to a limited volume or a shielded compartment by keeping all noise-free wires away from the pulse-carrying circuit. Wires placed close to the pulse circuit must be shielded or filtered. Putting a noise-free wire in a cable with even as little as a few inches of "hot" wire results in pickup of noise. The chassis should not be made a part of a high-current pulse circuit because the ground current is likely to induce noise in circuits that would otherwise be noise free. Instead, the pulse circuit should be completed in a coaxial line or a wire loop having the smallest possible area consistent with permissible capacitance, and grounded to the chassis at only one point. Care should be taken to see that currents are not induced across joints in shields or chassis or in noise-free wires by this pulse loop.

Obviously, care should be taken to prevent corona discharge and sparking in high-voltage d-c circuits. Full use should be made of the isolation properties of circuit elements, such as an electrostatic shield in a transformer, the inductance of the series field of a d-c motor, the shielding action of grounded leaves in relay spring piles, and so on.

*Shielding.*—For levels of internal noise less than 150 μv as measured with a 1-in. electrostatic probe, an enclosure should be complete except for louvers, with all mating surfaces paint free and bolted or spring-clipped together at intervals of a few inches. If the noise is greater than 150 μv, all openings should be screened and all joints should have continuous high-pressure contact by means of multifinger spring contacts or electric gaskets. These requirements are sometimes difficult to meet in pressure-tight containers. The sealing gaskets must not be placed between mating surfaces but should be pressed against the edges or backs of mating parts after electrical contact is made. When a flexible ground is required, a strap is to be preferred to a braid.[1]

*Filtering.*—Commerical noise filters having a wide range of attenuation characteristics are available. When the noise frequencies that must be attenuated are known, a filter can usually be found to serve. It is important however, that the filter be properly installed, preferably with one terminal projecting through the shield wall in the form of a bushing, and with the filter case firmly grounded to the shield wall. In fact, the feed-through capacitors of the bushing type, with a low series inductance, are excellent filters for higher frequencies especially when a resistor can be used in series with the lead inside the shield to form an *RC*-filter.

It has been found that for the noise frequencies generated by radar modulators, single *L*-section low-pass filters are adequate. In many cases they are superior to Π-sections because of the difficulty of grounding the Π-capacitors without a common impedance. Such filters, when designed by the simple theory which assumes pure lumped elements, should have a 10-db safety factor at the critical frequency. Single *L*-section filters designed in this way give 40 to 50 db attenuation from 0.2 to 20 Mc/sec and, sometimes, with excellent grounding and shielding, as high as 70 db. For attenuations of 60 to 100 db, two-section filters must be used. A line filter that has been found to be very useful consists of a 60-μh inductor with a 1-μf capacitor.

In using *L*-section *LC* filters, it is preferable to connect the capacitor to the noise source. With the reverse connection, it is sometimes possible for the series *LC* circuit to resonate and actually increase the noise voltage across the condenser.

Noise sometimes escapes from the beacon cabinet by way of the antenna itself. This can be minimized by use of a filter in the antenna circuit which narrows its pass band to just the amount required for the received and transmitted pulses. This can be done by using tuned stubs as filters, by using waveguide, or by making the antenna itself appear as a short circuit at the noise frequencies. For ungrounded open-wire transmission lines, in addition to stubs for short-circuiting off-frequency noise components, inductive coupling with a Faraday shield is desirable to prevent the line from acting as an antenna against ground at noise frequencies.

*Cabling.*—External pulse cables can cause much trouble by poor contact of the outer conductor at connectors. In general, double-braid

---

[1] A good discussion of shielding principles can be found in E. E. Zepler, *Techniques of Radio Design*, Wiley, New York, 1943, Chap. 8.

shielded cable should be used, with well-soldered coupling of the braid to the connector or a reliable mechanical coupling of high quality. A common mistake is to connect a shielded wire to a shielded pulse circuit or receiver by passing the wire through a grommet and grounding the braid to a point inside the shield. This makes the shield of the cable become an excellent inductively coupled antenna. The correct technique is to ground the braid completely to the outside of the shield to prevent this.

Flexible cables carrying high-power pulses often emit too much leakage radiation and must, therefore, be run through rigid metal conduits for shielding.

*General.*—The suggestions given above, though general, can be applied, by trail-and-error methods and experience, to a successful elimination of the radio interference caused by beacons. An example that shows the possible effectiveness of good circuit design and shielding, is a 200-kw pulse power modulator designed by the modulator group of the MIT Radiation Laboratory. It gave less than 50 μv of conducted noise from 0.15 to 20 Mc/sec *without* the use of filters.

## PROTECTION OF BEACONS FROM INTERFERENCE

**16·9. Noise Triggering.**—Spurious interrogating noise must be kept out of beacon receivers because short-pulse or "spike" interference, which would be only inconvenient in a communication receiver, can easily absorb a large fraction of the available duty ratio of a beacon.

There are two main types of interference to consider. The first is that caused by r-f signals in an adjacent channel received through the antenna (see Sec. 4·11). The second is caused by noise that enters the beacon by way of its antenna, power input or other cables, or is generated within the beacon itself.

Two things can be done to reduce the effects of interference of the second kind. First, the receiver circuit itself can be made as insensitive to noise as possible. The crystal-video receiver has a broad r-f pass band and is also very sensitive to video-frequency interference. The amplifier should be designed to have good low-frequency rejection to prevent amplification of microphonics and ripple from the power-supply. It should not have a video pass band that is wider than necessary for the required reproduction of pulse shape. The broadband superheterodyne beacon receiver, though approximately one hundred times more sensitive than the crystal-video receiver, does not present very much more difficulty because the high sensitivity is confined to a relatively narrow band.

Second, the interference must be kept out of the receiver, either by shielding the filtering the receiver completely or by attenuating the noise sources by shielding and filtering. In general, it has been found necessary to shield and filter both the receiver and the noise sources that cannot be eliminated. Methods of filtering and shielding the antenna circuit are discussed in Sec. 16·8.

Adequate shielding and filtering of noise sources within the beacon is sometimes very difficult. High-voltage corona, commutators, slip rings, vibrators, and, to a lesser extent, gas-filled and vapor-filled switch or rectifier tubes should, if possible, be avoided in the original design.

Fortunately, the modulator, the largest single source of noise within a beacon, does not cause much trouble because it operates only after the receiver has been interrogated, and the receiver usually is given time to recover by the blanking gate. If large video or r-f signals reach the various receiver grids, bias blocking may result from grid currents. However, normal shielding of the receiver is usually sufficient to prevent such blocking. The shielding of the receiver from the transmitter does not have to be as good in a beacon as in a radar set because the beacon receiver does not have to receive within a few microseconds after the modulator has been triggered.

A sufficiently large video or r-f signal from the first pulse of the transmitter can interfere with the proper formation of the remaining pulses by the coder. The coder, however, is relatively insensitive and little trouble is to be expected. If normal precautions are taken to confine the transmitter video and r-f energy, shielding of the coder will not be necessary—except, perhaps, to confine video noise generated by the coder.

If noise sources like d-c motors or vibrators are present in the beacon, however, effective shielding and filtering of the receiver may be necessary. For example, an airborne beacon was triggered by noise from d-c blower motors, in spite of filtering of the motors and grounding of a shielded i-f lead from the crystal mixer to the amplifier. It was finally found that this shield was improperly grounded. Changing the grounding of the i-f lead eliminated the trouble completely, after all attempts to filter the motors had failed.

A device that greatly facilitates the location of leaks in shielded receivers is a concentrated noise source. Such a source can be made by enclosing a battery-operated high-frequency buzzer in a very tight shield can, and bringing out a shielded coaxial lead from one of the buzzer contacts to a small exposed probe. Such a noise source can save many hours of trial-and-error work.

Even with a-c-operated beacons which have no moving contacts to generate noise, precautions must be taken to prevent noise from the power cables from getting into the receiver. With superheterodyne receivers, it is sufficient to shield the receiver and use a power-supply transformer that has an electrostatic shield around the primary winding. With crystal-video receivers filament chokes on the first video amplifier tubes are necessary to keep power-line noise from interrogating the beacon. In one early installation, video noise entering by way of the filaments of the crystal-video receiver was eliminated only by placing a double low-pass audio-video filter in the a-c line to the power supply of the receiver. Commercial noise filters were of no use because the offending noise was of lower frequency than they could reject.

Battery-operated beacons with vibrator or dynamotor supplies present difficult problems of protecting the receiver because both dynamotors and vibrators are strong sources of noise over a great range of frequencies. It is necessary to shield and filter vibrator supplies completely, and sometimes necessary to shield dynamotors, although the dynamotor case itself is usually sufficient shield if the leads are filtered very close to their point of exit from the dynamotor.

With battery-operated beacons having vibrator supplies, it is not only necessary to shield and filter the d-c outputs of the power supply very carefully, but the a-c voltage generated in the internal impedance of the battery by the pulsing load of the vibrator must also be filtered out of the filaments of the receiver tubes. This is particularly important with crystal-video receivers using filament type tubes. The filtering is usually done with a Π-filter having an iron-core inductor of several henrys inductance and two electrolytic capacitors of 25 $\mu$f or more capacitance (see Fig. 15·8).

The brushes of dynamotors to be used at high altitudes present a special problem. Ordinary brushes which do not spark at sea level spark badly at high altitudes and cause intense triggering of a beacon. The life of the brushes is short under these conditions. Satisfactory brushes for use at high altitudes are now available, but even when they are used, the shielding and filtering of the dynamotor should be good in order to allow for the increased sparking that will result from reduced pressures and wear of the commutator.

In sum, the problem of radio interference in beacons is solved by skillful application of the principles of circuit design, shielding, and filtering. It must be admitted, however, that application of these principles is an art and not an exact science, so that methods of trial and error are often necessary.

**16·10. Suppression.**[1]—It is often necessary to install a beacon so close to other radar equipment or other beacons that continuous triggering of the beacon when the neighboring set is in operation cannot be prevented. When such triggering occurs, it can be prevented only by using a signal from the interfering set to suppress the response of the beacon by making it insensitive for a short interval. The suppression trigger is used to generate a suppressor gate. If suppression is not used, the needless

---

[1] Secs. 16·10 and 16·11 by J. J. G. McCue and M. J. Cohen.

responses will waste an appreciable part of the available duty ratio of the beacon, and will clutter the screens of the interrogators that are actually using the beacon.

The signal which suppresses the beacon may be transmitted to the beacon either as a video pulse on a wire, or as an r-f pulse through space. Clearly, the suppression signal must act on the beacon before the interfering signal produces the undesired response. Since the beacon receiver introduces some delay, a suppressor gate triggered by the suppression signal can usually be made to render the coder inoperative by the time the radar signal has passed through the receiver. If the suppression signal is generated before the radar signal, the problem is simplified. In most cases, the suppression signal from the radar will be used merely as a trigger for generating the suppressor pulse. This pulse may be generated by either a multivibrator or a blocking oscillator.

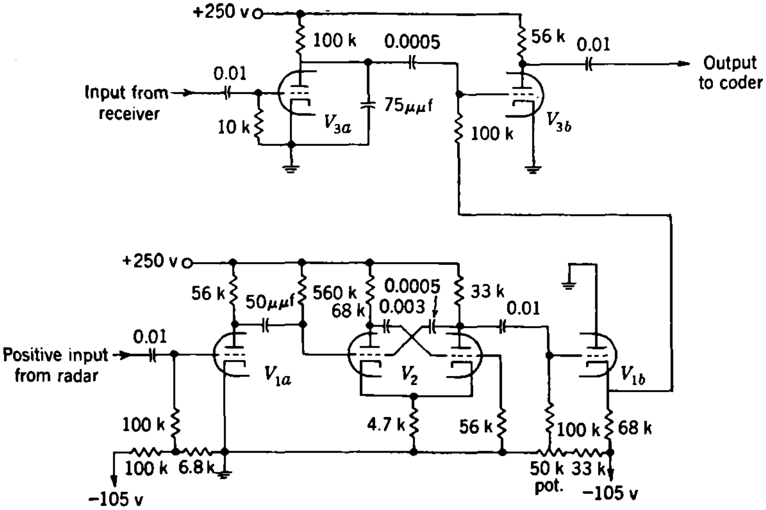


FIG. 16·1.—A suppressor circuit to prevent interrogation by a near-by radar. NOTE: All tubes are 7F8's.

The suppressor pulse must satisfy the following conditions: (1) it must make the beacon inoperative before the interfering signal triggers it; (2) it must not itself trigger the beacon through stray capacitance; (3) it must not interfere with normal beacon operation if the beacon has already been triggered by a desired signal just before the suppressor signal arrives.

The suppressor pulse can be used to render inactive the tube which triggers the coder. Figure 16·1 shows a highly satisfactory circuit, which introduces an adjustable delay and incidentally acts as a pulse-width discriminator. If the receiver already introduces enough delay, or if the video suppression trigger from the radar is generated before the r-f pulse, the delay circuit in Fig. 16·1 is unnecessary. In this case $V_3$ could be omitted, and the pulse from $V_{1b}$ could cut off a tube in the video amplifier of the receiver.

When a video connection between the beacon and the interfering radar is impractical, r-f pickup may be used. This system is useful, for example, in ground beacons that are close to a maintenance base for airborne radars. It involves the use of an auxiliary receiver of low sensitivity which will detect signals from radar sets up to about 1 mile away. The auxiliary receiver generates the suppression signal for the beacon. The antenna for the auxiliary receiver should have a pattern covering only the area from which the disturbing signals are emitted. This will minimize the unintentional suppression of responses to radars which are trying to interrogate the beacon from close range. The required sensitivity of the receiver depends, of course, on the antenna used. The receiver may be of the crystal-video type, with a sensitivity of the order of 40 db below that of the beacon receiver. Unless the beacon contains a decoder that delays the signal received by the beacon receiver by an appreciable amount, the video bandwidth of the auxiliary receiver must be much wider than that of the beacon receiver. This is necessary for the suppressor gate to be generated soon enough, and is feasible because of the lower gain of the auxiliary receiver. The output of the receiver is a video pulse which can be used as a suppression trigger in the manner described above.

The r-f transmission of suppression signals has not been used with airborne sets. An airborne beacon will usually be interfered with only by a radar in the same plane. When this is likely to happen, video transmission of the suppressor pulse saves space and weight. Often, however, a radar will not interfere with a beacon in the same aircraft, particularly if the radar and beacon frequencies are very different.

Radio-frequency filters are sometimes useful to prevent interference. For example, a 10-cm beacon with a crystal-video receiver can be protected from a 3-cm radar in the same plane by installing a 3-cm r-f rejection filter in the beacon receiver. (See Sec. 7·11.)

A beacon with a superregenerative receiver sometimes radiates enough noise from the quenched oscillations to jam a near-by interrogator receiver operating on the same frequency. The superregenerative receiver can then be suppressed, by a video pulse, for the duration of the range sweep in the interrogator. This is a drastic remedy, because it puts the beacon out of action for a large fraction of the time. This constitutes an argument against using superregenerative receivers in beacons operating on the same frequency as a near-by interrogator.

**16·11. Crystal Protection.**—Crystal detectors in microwave beacons often are damaged by pulse powers of the order of 1 watt, and must, therefore, be protected against being burned out or damaged by their own transmitters or by others near by.

The crystal in a beacon using a single antenna is protected by the duplexing system. Here the transmitter fires a gas-filled TR-switch tube which places a protective short circuit in the line to the crystal. Wideband duplexers of current design function only on high-power pulses, and are, therefore, not suitable for low-power beacons.

If the beacon uses two antennas, they must be so disposed that the leakage from one antenna to another is low enough to prevent crystal damage. In a portable or airborne set with removable antennas, the r-f leakage may increase to an unsafe level when the antennas are removed. This is especially true of 3-cm beacons using waveguide parts that are adjacent to each other. A short-circuiting post, automatically inserted into the waveguide when the antennas are removed, will protect the crystal by reflecting power that leaks into the guide. In a beacon that operates intermittently, crystal damage can be reduced by installing in the receiver r-f line a short-circuiting device called a "crystal gate" that is withdrawn automatically (by a solenoid) when the beacon is turned on, but falls back into its protecting position when the beacon is turned off.

The use of filters to protect crystals from damage due to neighboring systems is discussed in Sec. 7·11.

## PACKAGING

BY B. W. PIKE

Packaging of the major components into a beacon is difficult because none of the numerous configurations possible is ever entirely free of faults. A careful study of the many possibilities, however, will yield dividends. The packaging of any beacon must take into account certain general considerations as well as considerations special to the unique purposes of the beacon.

**16·12. General Considerations.**—The first of these are considerations of weight, size, and shape, which may be general enough so that the designer can give most consideration to other factors in deciding on the final shape. Sometimes, however, the intended use of a beacon is so specific that rigid specifications for weight, size, and shape must be laid down.

Another important aspect of package design is shielding. Beacons are usually packaged in metal to assure adequate shielding as well as adequate mechanical strength. One recent design uses a metal-lined plywood case.

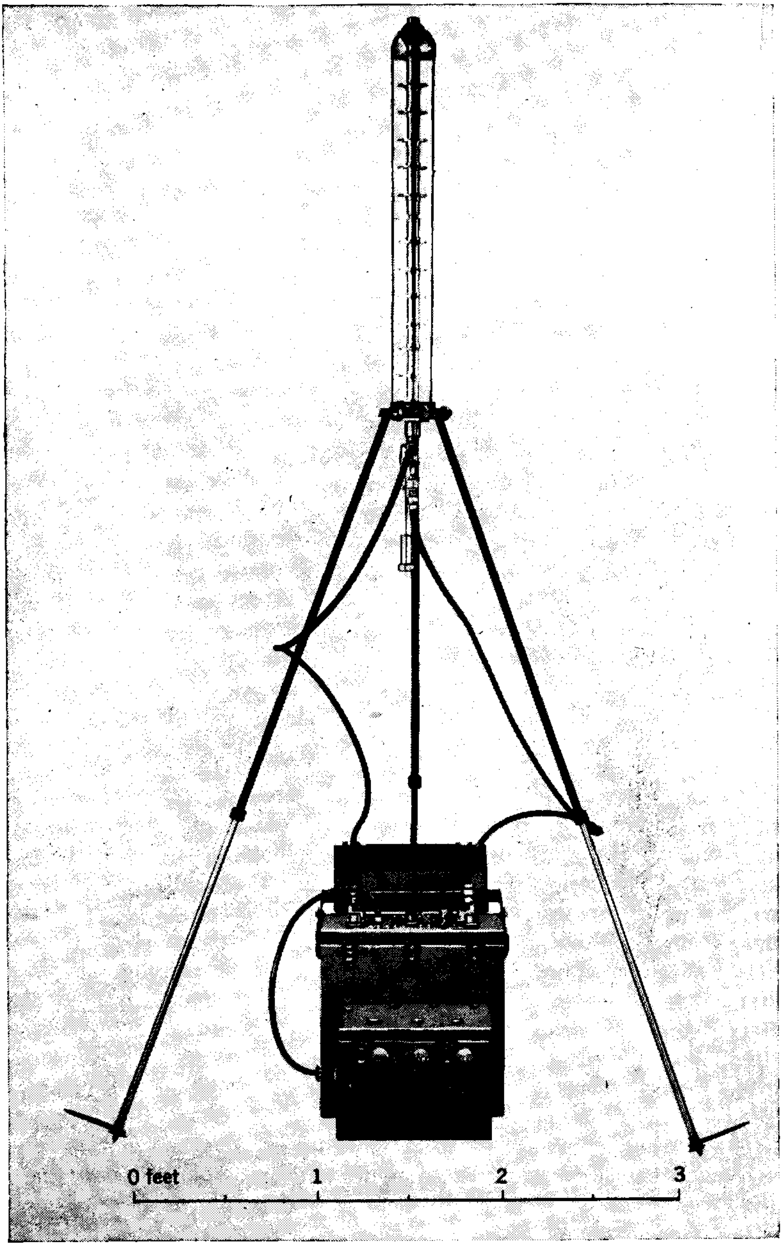


FIG. 16·2.—AN/UPN-1, a lightweight 10-cm battery-operated beacon with waterproof packaging.

*Weatherproofing and Heat Dissipation.*—Weatherproofing and heat dissipation are important but difficult aspects of beacon packaging. Weatherproofing includes "splashproof," "waterproof," and "pressurized" packaging. In splashproof packaging, the components themselves are made moistureproof. This packaging permits free access of moisture-laden air but shields the components from direct rain or splashed water. This type of packaging has the advantages of simplicity and of offering a simple solution to problems of dissipation of heat; it is the type most widely used for large and medium ground beacons. Splashproof packaging is illustrated by the large 3-cm ground and shipboard beacon which is shown in the photograph, Fig. 15·11, Sec. 15·12.

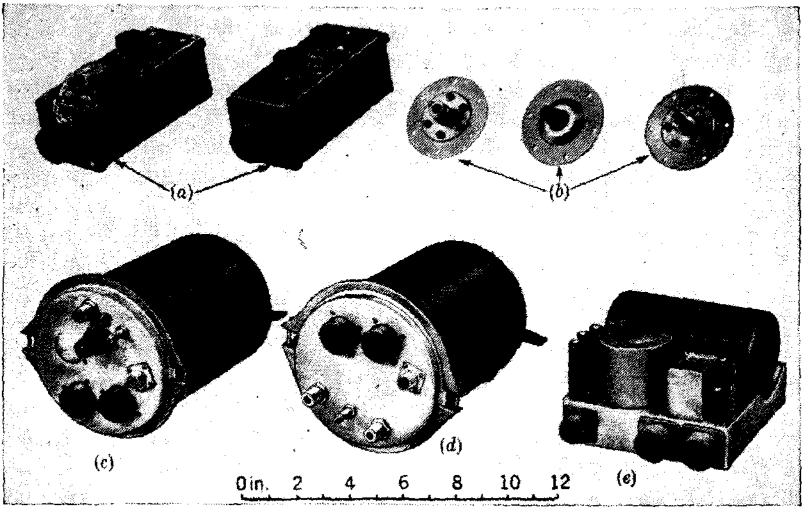


FIG. 16·3.—AN/APN-19, an airborne beacon illustrating pressure-type packaging. (*a*) Control boxes. (*b*) Antennas. (*c*) Transmitter. (*d*) Receiver. (*e*) Dynamotor.

Where the amount of heat to be dissipated permits, the preferred packaging is the waterproof type, in which the beacon is hermetically sealed and a dehydrating agent is placed within the cabinet. This type of packaging allows the use of ordinary nonwaterproof components, which are smaller and lighter than those which have been waterproofed. It is illustrated by a lightweight 10-cm battery beacon, AN/UPN-1, the prototype of which is shown in Fig. 16·2.

The third type, the high-altitude hermetic packaging, differs from the waterproof type only in that the container must be capable of maintaining an appreciable pressure differential. For much airborne equipment intended for high-altitude use, this type of packaging is essential, even when the problem of heat dissipation complicates the design. This type of packaging is illustrated by AN/APN-19, an airborne 10-cm beacon, shown in Fig. 16·3.

**16·13. Vibration and Shock Mounting.**—Vibration and shock must be considered in the packaging of any beacon. Sometimes the beacon is made rugged enough to operate without special mountings, but usually it is mounted on rubber. The subject of shock and vibration mounting is complex. A beacon designer, therefore, will do well to consult with experts. Commercially available rubber shock mountings do a fairly satisfactory job if the designer chooses the correct stiffness of disk and places the disk correctly. In general, a stiff mounting is desired for protection from shock and a soft mounting is required for vibration. The lower the vibration frequency, the softer the mounting must be. If possible, the disks should be placed symmetrically around the center of gravity and should be loaded in shear. In order to accommodate vibration or shock in all three planes, two or three sets of ordinary disks are needed. Recently developed special mountings using sponge rubber are very effective, but each must be custom-designed for a given device.

**16·14. Other Considerations.**—Miscellaneous general considerations, like ease of operation, ease of maintenance, and safety must be considered in packaging design. Ease of operation can be achieved by having all the required meters, controls, and indicators conveniently placed for manipulation and inspection.

Ease of maintenance is more difficult to achieve. A compromise usually must be made between ease of maintenance and the extra space and weight required to give easy access to all parts of the beacon. A method which is always used for large beacons is to install the different major components as separate removable units. The same principle can well be applied to small beacons in order to waste no volume in the form of access space.

For safety from high voltages, it is advisable to use door interlocks, warning signs, pilot lights, and guards. Bleeders, or better, automatic door-operated grounding switches, should be used on high-voltage supplies. When blowers are used to ventilate cabinets, the blower should draw air through a filter in order to keep dust out of the beacon.

Beacons intended for specific uses usually have a few special packaging requirements. These will be discussed by listing various uses of beacons with some examples.

*Ground and Shipboard Beacons.*—Such beacons are usually large, complex, high-performance sets. For convenience in shipboard installation, it is preferable to have the beacon require access from the front only. Also, the height should not be greater than about 6 ft. For convenience in handling, the beacon should have hoisting eyes on the top and several handles on the sides.

For ship installation, vibration mounting is difficult because of the low frequencies involved. On warships, this problem is complicated by the very high shock of gunfire. In such cases a dual mounting of stiff and flexible mounts in series is usually necessary. AN/CPN-6, shown in the photograph, Fig. 15·11, of Sec. 15·12, is a large 3-cm ship and ground beacon which illustrates the typical packaging for such a beacon. It is housed in two relay-rack cabinets, with the major components in drawers. The transmitter, modulator, and high-voltage supply are in one cabinet. The other cabinet contains the receiver, coder, and driver. Each cabinet is equipped with thermoswitch-controlled blowers.

*Lightweight Beacons.*—Beacons for portable use should be light and easy to carry. The weight of any single package should be no greater than about 50 lb. The shape should be such that the beacon can be carried easily by hand or as a back pack. The beacon should also be easy to assemble and place in operation, and because shelter is not always available, a portable beacon should have hermetically sealed waterproof packaging. Also, because of the bumps and blows that hand-transported equipment receives, the beacon should be either internally shock-mounted or carried in a padded case.

An example of this type of packaging is the 10-cm battery-operated beacon, AN/UPN-1, shown in Fig. 16·2. The beacon proper weighs 35 lb, is internally shock-mounted, and is of waterproof packaging. The antenna, also waterproof, weighs 15 lb. The storage battery box weighs 25 lb.

*Airborne Beacons.*—Airborne beacons must be packaged to obtain the lowest possible weight and volume. Although pressure-type packaging is usually necessary, it requires extra weight and should be avoided if the beacon can be designed to operate satisfactorily without it. When the beacon is being designed to fit aircraft in which space will be allotted to it, the beacon should be built into a single container and have a single control box built into the aircraft radio control panel. However, if the beacon must fit into an aircraft already crowded with equipment, it may be preferable to package it in several units which can be placed in whatever space is free.

A 10-cm airborne beacon, AN/APN-19, is shown in Fig. 16·3. This beacon was designed to fit into an already crowded aircraft. The receiver and transmitter are each in a pressure-tight cylindrical container. The remaining parts of the beacon are the dynamotor power supply, two control boxes, an r-f filter, and three antennas. This beacon weighs 35 to 40 lb installed.

## PART III

# INTERROGATOR AND SYSTEM DESIGN

In the following chapters the design of interrogators and examples of complete systems are treated.

Interrogator design is discussed briefly. Only features of radar sets concerned with interrogation of beacons are taken up; radar design itself is covered in other volumes of this series. Design of interrogator-responsors is similarly treated; only those special features which differ from radar design are discussed.

# CHAPTER 17

# RADAR DESIGN FOR BEACON OPERATION

By W. M. Preston

It has already been emphasized in this volume that a radar beacon and the equipment which is to interrogate it form a system and should be designed together. This is obviously true in the case of an interrogator-responsor unit, the sole purpose of which is to work with beacons. The primary function of a radar set—be it sea search, fire control, early warning, or any other—will usually take precedence in its design. It is precisely because compromise may be necessary that the secondary beacon function should be planned in advance, and its design correlated with the design of the beacon itself. It seems useful, therefore, in a book dealing with beacons, to discuss those parameters of radar equipment that concern satisfactory operation with beacons. Since radars now being designed are predominantly microwave sets, we will be considering primarily microwave radar-beacon systems.

**17·1. The Transmitter.** *Power Output Control.*—One radar parameter usually determined by echo considerations is the power output of the transmitter. With the antenna gain, it determines the maximum range of the radar with a given beacon, as discussed in Chap. 2. The power output also influences the minimum range at which the beacon will be triggered on the side lobes of the antenna pattern, with resulting loss in azimuth discrimination (see Sec. 2·11) and possible overloading of the beacon.

The minimum range at which side-lobe triggering occurs could be decreased if the operator could reduce the power put out by the radar transmitter at will. Continuously variable power dividers have been made, but they are relatively complicated and require the operator to manipulate an additional control. A compromise to be considered is a fixed r-f attenuator, with an attenuation of about 20 db, to be introduced into the transmitter line at the discretion of the radar operator by a switch on the control panel.

In a plane homing on a beacon, the fixed r-f attenuator would be used as soon as side-lobe triggering began, and it would reduce the range at which it occurs by a factor of 10. The value of 20 db is suggested because it is approximately the ratio between the power in the main lobe and that in the first few side lobes of a good antenna. If, at a particular range, the power in the side lobe is just sufficient to trigger a beacon, the main-lobe intensity reduced by 20 db should be sufficient to trigger it.

*Frequency Stability and Scatter Band.*—Frequency stability[1] in the radar transmitter is not critical when operation over a wide scatter band is allowed. The beacon receiver must cover the entire band; it can usually be made somewhat broader to allow for possible frequency drift in the radar transmitter.

In some systems, however, operation on assigned spot frequencies is required in order to reduce interference. The radar must then have a tunable transmitter and provisions must be made in the field to tune it to the correct frequency channel. It is desirable to make the pass band of the beacon receiver as narrow as possible in order to increase selectivity, but before choosing a value for its bandwidth, specifications for the frequency stability of the radar transmitter must be determined. This requires a study of variations of frequency with temperature, line voltage, and other factors, and of the tolerances that can be maintained in the test equipment which is used in adjusting the transmitter. Similar tolerances are necessary for the beacon, and all must be considered together in deciding on the pass band of the beacon receiver. Few decisions are more difficult to make in the early stages of design of a system because of the large number of interrelated variables.

*Width and Form of Transmitted Pulse.*—The conditions on the transmitter pulse are not stringent in systems that do not employ pulse-width discrimination. Even when spot frequency channels are used, the bandwidths of beacon receivers are normally greater than the bandwidths required to pass radar pulses of normal duration because of the tolerances discussed above. Systems demanding maximum range accuracy must use transmitter pulses with steep leading edges. A beacon is triggered when the incoming pulse reaches a certain required amplitude; if the leading edge of the pulse slopes, the time at which the beacon is triggered will vary with strength of the signal and there will be a corresponding variation in the apparent range to the beacon.

In contrast, pulse-width discrimination requires that strict tolerances be set on the duration of pulses from the radar transmitter. As an illustration, consider microwave ground beacons for aircraft navigation and the airborne radars which operate with them. The durations of search pulses are limited to values less than 1.0 μsec or greater than 5.0 μsec, while beacon interrogating pulses must be between 2.0 and 3.0 μsec. If it is decided that a tolerance of 10 per cent is the greatest that can be held in the field, radar pulse widths should be designed for 0.9 μsec or less for search and for 2.2 μsec for beacon interrogation. Similarly, to allow for setting errors, the beacon discriminator should be set at 1.9 μsec.

---

[1] General considerations of frequency choice are covered in Chap. 4.

*Multiple-pulse Interrogation Coding.*—Multiple-pulse interrogation coding is introduced into some systems in order to obtain interference-free channels; the radar modulator must be designed accordingly. For example, a double-pulsed system which employs an airborne beacon and a modified SCR-584 has been used. The latter has a crystal-controlled timing circuit which supplies timing pips at intervals of approximately 12 μsec; this rate is counted down to give the pulse-repetition frequency, 1707 cps. To obtain an 8-μsec code space, each 1707-cps pulse triggers a phantastron delay of 4 μsec. The phantastron output then triggers the modulator, forming the first code pip. A gating circuit selects the next timing pulse, which occurs 8 μsec later, permits it to trigger the modulator a second time and, simultaneously, to start the range sweep. Now, the beacon cannot reply to a double-pulse interrogation until the second pulse arrives, but if the start of the radar range sweep is delayed until the second transmitter pulse, the beacon signal will appear on the indicator at the proper range. The problem of the tolerances to be placed on the spacing of the pulses in multiple-pulse interrogation systems has been discussed briefly in Sec. 5·5.

*Pulse-repetition Frequency of the Transmitter.*—The minimum detectable signal on the radar indicator decreases inversely with the square root of the pulse-repetition frequency, as discussed in Sec. 2·6. This indicates the desirability of high rates of interrogation. Even if an upper limit on the pulse-repetition frequency is not dictated by the range sweep used, such a limit must be set arbitrarily to prevent overloading the beacon transmitter if high traffic-handling capacity is required. Pulse-repetition frequencies greater than about 500 cps are rarely used for beacon operation; values as low as 200 cps are often satisfactory.

## THE RECEIVER

**17·2. Bandwidth.**—The i-f bandwidth of the radar receiver is determined by the relation $B = a/\tau$, where $B$ is the bandwidth in megacycles per second, $\tau$ is the beacon transmitter pulse duration in microseconds, and $a$ is a numerical factor which has the theoretical value of 1.2 for best signal-to-noise ratio, but which in practice is found to be better taken as 1.5.

If radar search functions alone determine the i-f bandwidth, the beacon pulse width will necessarily be chosen accordingly. When this was done in the past, for example, ground beacons for aircraft navigation were limited to minimum pulse widths of 0.5 μsec because of the small bandwidths of conventional radar receivers. From the point of view of beacon design, shorter pulses are sometimes desirable in order to decrease the duty ratio and thus to increase traffic capacity. It is advantageous, therefore, to have a broader i-f pass band whenever the slight reduction in receiver sensitivity entailed can be tolerated.

Radar receivers have been made with variable i-f bandwidth; a narrow band could be used for search and a broader one for beacon reception. A still more flexible arrangement is that of using an entirely separate receiver for beacon reception. It will be described below.

The beacon usually replies on a nominal spot frequency. The local oscillator in the radar receiver is frequently tuned for reception at this nominal frequency by an AFC circuit. Whether AFC is used or not, it is necessary to allow a tolerance. There is, therefore, a strong reason to make the i-f bandwidth of the radar receiver somewhat greater than the value determined by the width of the beacon pulse alone, so that optimum reception will be obtained with due allowance for the tolerances on beacon frequency and on tuning of the local oscillator of the radar.

*R-f bandwidth.*—Until recently, duplexing in microwave radars was accomplished by means of gas-filled discharge tubes in relatively high-$Q$ resonant cavities tuned to the radar frequency. The loaded $Q$'s are between 100 and 300. The earlier airborne radar sets thus suffered a loss up to 20 or 30 db in receiving beacons; in the worst case, the beacon reply could differ by as much as 2 per cent from the radar frequency to which the TR switch is tuned. This necessitated a corresponding increase in the power output of the beacon and made it impossible to balance the interrogating and receiving links over the whole scatter band. The first solution introduced was that of retuning the TR cavity. A magnetically controlled slug was adjusted so that the TR cavity was tuned to the radar frequency when the slug was inserted, and to the beacon frequency when it was withdrawn. The slug was actuated automatically by the search-beacon switch.

More recently, TR and ATR tubes which cover a 2 per cent frequency band (or more) with a signal loss received of less than 2 db have become available (see Sec. 7·9). These tubes make it possible to spread radar transmitter frequencies over a corresponding scatter band without any need for retuning the TR and ATR cavities. At the same time, they eliminate most of the loss in beacon reception.

**17·3. Receiver Channeling.** *Single-receiver Channel.*—If the reply frequency of the beacon differs from that of the radar transmitter and a single receiver is used for both, either the local oscillator must be retuned or two local oscillators must be used. Because the scatter bands of airborne radar transmitters at both 10 and 3 cm are broader than the electrical tuning range of local oscillator tubes now available, the second alternative has usually been chosen. The two local oscillators usually feed into a common mixer; it is then impossible to run both continuously because of pickup of high harmonic beats falling within the i-f band of the receiver. The heaters alone are left on and the plate voltage is switched from the search local oscillator to the beacon local oscillator by the search-beacon switch. Recent r-f circuit developments, including use of separate mixers, have made it possible to operate both local oscillators simultaneously in many cases.

Precise tuning of the local oscillator for beacon signals may be accomplished in several ways. A manual tuning control which relies on the reception of a beacon signal as a tuning indicator can be provided. Alternatively, some of the local-oscillator power may be led through a high-$Q$ standard cavity to a crystal rectifier and meter output; maximum meter deflection then indicates when the local oscillator is correctly tuned. It is then unnecessary to hunt for a beacon signal while tuning manually. Most satisfactory of all is the use of an AFC circuit, which tunes the local oscillator automatically so as to "peak" the standard cavity output. "Push-button" beacon reception, which will almost certainly be demanded in future systems, is thus made possible. Details of beacon AFC circuits can be found in *Microwave Receivers,* Vol. 23, Chap. 3, Radiation Laboratory Series.

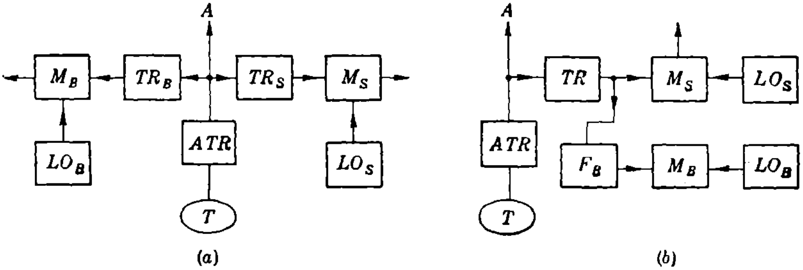


FIG. 17·1.—Schematic representation of the r-f components in two types of dual-channel operation. (*a*) Separate beacon and search TR cavities are used; they may be high-$Q$. (*b*) A single broadband TR cavity is used with an auxiliary filter for the beacon frequency.

*Separate Receiver Channels.*—The method of reception of off-frequency beacon response that is most flexible provides separate receiving channels for search and beacon signals. This implies, in general, separate TR tubes, mixers, local oscillators, detectors, and i-f amplifiers. The signals can be mixed after the second detector and fed into a common video amplifier or, preferably, they are mixed after some video amplification. The following advantages are realized.

1. The i-f bandwidth of the beacon receiver can be chosen for optimum reception of beacons without having to meet search requirements.
2. Separate time-varied gain circuits can be provided for search and beacon (see Sec. 17·4).

. Separate controls of i-f gain and video output level can be used; beacon and search signals can be presented with different intensities (see Sec. 7·4).

Figure 17·1 illustrates two arrangements of the r-f components which ovide separate receiving channels. In Fig. 17·1*a* the two r-f channels e completely separate. If the beacon and search functions are on dif-

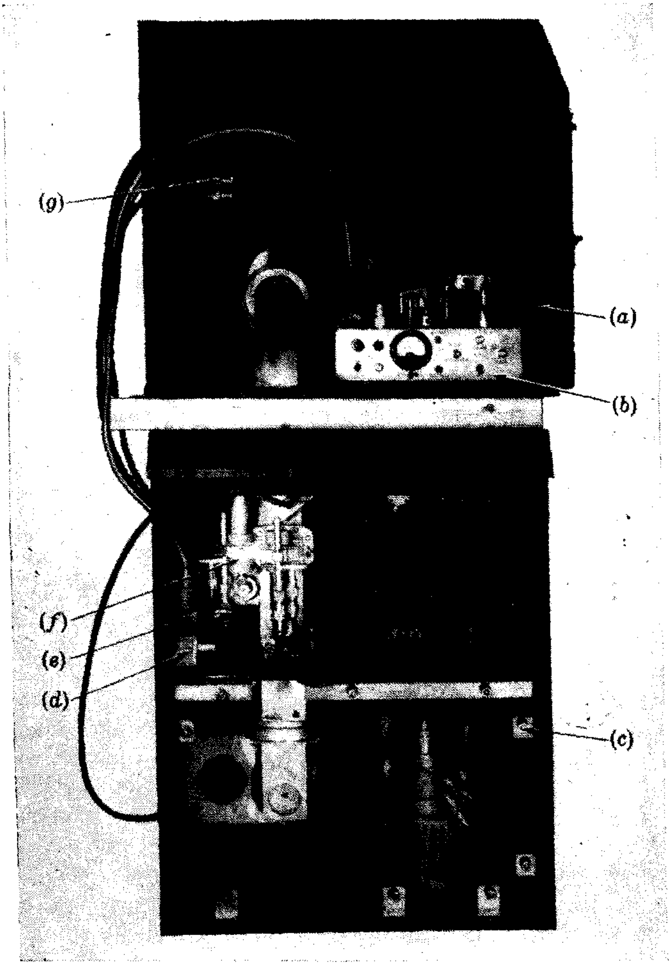


FIG. 17·2.—Beacon receiver components of a ground radar set, exclusive of displ (*a*) Beacon box. (*b*) Main chassis. (*c*) VL box. (*d*) I-f strip. (*e*) Junction l (*f*) Beacon TR cavity. (*g*) Connector well.

ferent spot frequencies, the two TR cavities can have relatively high $Q$'s, each will offer a high impedance at the frequency to which the other is tuned, and there will be negligible loss on either channel resulting from the presence of the other. If the search function requires a broad scatter band, the search TR cavity will be of the low-$Q$ type and will present a low impedance at the beacon frequency. As a result, beacon reception will suffer a loss of approximately 3 db, half of the signal being lost in the search receiver. Figure 17·1*b* shows an alternative arrangement which employs a single broadband TR tube, and a relatively high-$Q$ filter in the beacon line. Its advantage over the previous circuit is that a filter cavity requires less attention than a TR cavity.

Figure 17·2 shows a photograph of the beacon receiver components of a 10-cm ground radar system. These are all the components, exclusive of the displays, which are required for adding beacon facilities to the radar set. Some of the components shown are associated with the transmitter and receiver for the vertical lower (VL) beam of the system (cf. Chap. 19).

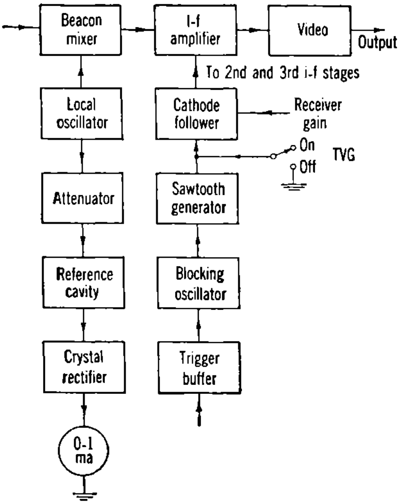


FIG. 17·3.—Block diagram of beacon receiver. Note wide i-f pass band (8 Mc/sec), adjustable time-varied-gain control, and independent video signal level control. AFC of local oscillator is not shown (see Fig. 8·14).

Figure 17·3 shows a block diagram of the complete receiver system. Noteworthy features are its 8-Mc/sec-wide i-f amplifier, the AFC of the local oscillator, and the time-varied control of the i-f gain. The last of these requires further discussion.

**17·4. Receiver Gain Control and Time-varied Gain.**—We have already mentioned, in Sec. 17·1, the loss of azimuth discrimination which can occur at short ranges because of the presence of side lobes in the radar antenna pattern. The ideal solution, as pointed out there, is to reduce the emitted power of the radar enough to prevent triggering the beacon by side lobes at short ranges. If this is not practical, azimuth discrimination may be obtained by reducing the gain of the i-f amplifier in the radar receiver. This has the drawback that frequent adjustment of the control is necessary for an airborne radar homing on a beacon. An automatic-gain-control circuit which varies the gain during each sweep partially overcomes this difficulty. This system is called "time-varied gain."[1] In TVG, the i-f gain is reduced at the start of each range sweep and allowed to recover to its normal value at a predetermined range. The ideal rate of change of gain would follow the inverse square of the range for beacon reception and the inverse fourth power of the range for radar search. The provision of separate receiving channels allows optimum independent adjustment of the TVG-circuit constants, while a compromise must be accepted if a single receiver is used simultaneously for both search and beacon.

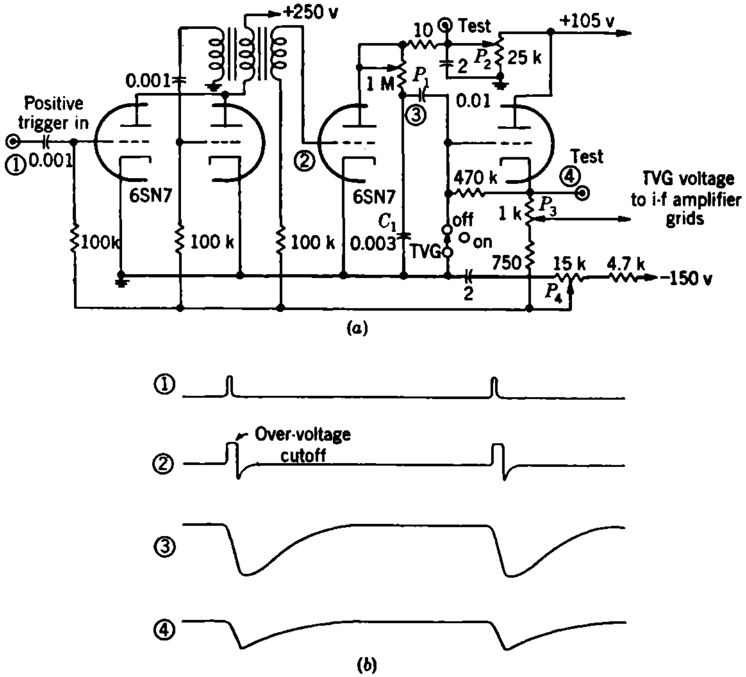


FIG. 17·4.—(a) Circuit diagram of beacon receiver time-varied-gain control circuit. (b) Waveforms at numbered points of (a). $P_1$—time-varied gain range control. $P_2$—TVG amplitude control. $P_3$—manual gain control. $P_4$—Maximum gain adjustment.

Figure 17·4 shows the time-varied gain circuit used in the beacon receiver of a ground radar system. The input trigger, which is coincident with the transmitted pulse, is amplified in the buffer and initiates regeneration in the blocking oscillator. The blocking oscillator, during its pulse, drives the grid of the sawtooth generator to conduction. Prior to this pulse the condenser $C_1$ is charged to the voltage determined by the

---

[1] In the past it has been widely known as "sensitivity-time-control," STC.

setting of $P_2$. During the pulse, $C_1$ is discharged rapidly through the sawtooth-generator tube. Immediately after the pulse, the voltage on $C_1$ begins to rise exponentially toward its former value with a time constant adjustable by means of $P_1$. The sawtooth waveform is passed through the cathode follower and applied as a bias voltage to the grids of the second and third i-f amplifiers in the receiver, where it controls the receiver gain.

With the TVG switch in the OFF position, the grid of the cathode follower is grounded, and the cathode is near ground. Potentiometer

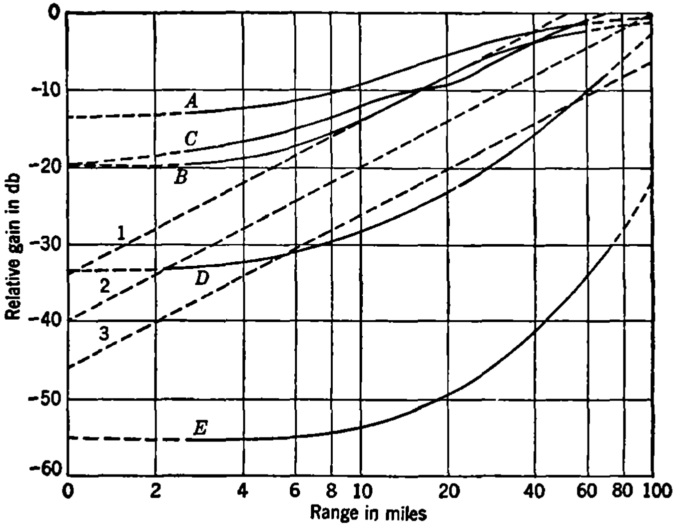


FIG. 17·5.—Time-varied-gain control characteristics.

$P_3$ provides manual control of the receiver gain by controlling the bias of the second and third i-f amplifiers. The maximum gain adjustment $P_4$ is adjusted so that the voltage at the cathode of the cathode follower is the same for both positions of the TVG switch; this results in equal maximum values of gain for both manual and automatic gain control. The TVG amplitude control $P_2$ adjusts the amount of change in sensitivity and $P_1$ and $P_2$ together control the time (or range) of complete recovery of sensitivity. Figure 17·5 shows the range-sensitivity characteristics that can be obtained with several settings of $P_1$ and $P_2$, as well as the theoretically desired curves to compensate for the inverse square of range.

The strength of the beacon signal received by a radar increases inversely with the square of the range only if the product of the effective antenna gains remains constant. At high altitude and short range, an airborne radar may receive only a weak signal from a ground or ship beacon, unless the beacon antenna has been specially designed to have good overhead coverage (see Secs. 7·3 and 7·4). Time-varied gain in this case reduces the gain most just when maximum gain is needed. Therefore, if good overhead coverage by an airborne radar is desired, it may be preferable to use a fixed attenuator of the type described in Sec. 17·1, in the r-f line. If the attenuator is placed in the r-f line between the antenna and the TR switch, it will reduce the strength of both the outgoing (transmitted) signal and of the incoming (received) signal. Like the manual gain control, however, its use requires attention on the part of the operator.

The performance of the TVG circuit will be less dependent on the characteristics of individual tubes if the TVG is used to vary the cathode current in the i-f amplifier stage rather than the grid bias.

## ANTENNA DESIGN

**17·5. General Properties.** *Beamwidth and Stabilization.*—Other factors being equal, the accuracy with which azimuth may be read is inversely proportional to the half-power width of the radar antenna beam. Antenna stabilization is a virtual necessity for narrow-beam shipborne radars and a great help for airborne radars with narrow beamwidths. In the latter case the added weight may be a serious factor.[1]

*Polarization.*—Beacon antennas can be made with either horizontal or vertical polarization. The choice can be made for any radar set, therefore, on the basis of optimum conditions for the radar function alone. As we have seen in Chap. 3, nulls give less trouble with vertical than with horizontal polarization.

However, all types of radar which are to operate with any one type of beacon must have the same direction of polarization, because cross polarization produces a large loss. Beacon antennas have been made with circular polarization, but such antennas necessarily involve a loss in signal strength of 3 db when used with plane-polarized radar sets. Vertically-polarized single-element beacon antennas for installation on aircraft are often considerably more compact than horizontally polarized antennas intended for aircraft use, so ground or shipborne radars for use with airborne beacons are preferably polarized vertically.

Certain automatic-following radars have used an off-axis rotating dipole as the antenna feed for lobe switching. This supplies continuous information for radar search because the cross section of most targets does not vary greatly with polarization. For linearly polarized beacon

---

[1] For a discussion of antenna stabilization, refer to *Radar Scanners and Radomes*, Vol. 26, Chaps. 4 and 7, Radiation Laboratory Series.

antennas, however, the signal strength varies as the square of the cosine of the angle between the plane of polarization of the beacon and that of the radar. A nutating eccentric antenna feed which maintains the plane of polarization constant has been developed and should be more satisfactory with beacons because it will give continuous interrogation and more nearly constant reply signals.

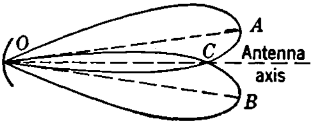


FIG. 17·6.—Crossover point of lobe-switched antenna. *A* and *B* are the directions of the main lobe of the antenna pattern in its two extreme positions. The length of a vector from *O* to any point on either curve is proportional to the power radiated in that direction. The crossover point *C* of the two lobes is, in this case, at about 75 per cent of maximum power in the directions *OA* or *OB*.

*Crossover Point.*—Antenna lobe switching is frequently employed in radars designed for getting accurate angular information. The crossover point of the two lobes must be selected for best operation, especially with automatic-tracking sets like the SCR-584 (see Fig. 17·6). Suppose, for example, that it is desirable to have the received signal strength from a target in the direction *OC* of the antenna axis 50 per cent of the received signal strength when the target lies along *OA* or *OB*. For tracking by radar echoes, the crossover point must then be set at 70.7 per cent. The target on the axis will then be illuminated with 70.7 per cent of the maximum energy, and the return signal will be reduced by 70.7 per cent, a total reduction of 50 per cent. When tracking a beacon, however, the crossover point must be set at 50 per cent. The strength of the beacon signal is independent of the strength of the interrogation pulse, as long as it is above the threshold for triggering; the lobe-switching discrimination enters only into the receiving link. It may be necessary, therefore, to use antennas with different crossover points for best operation with beacons and radar echoes.

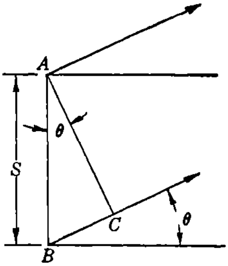


FIG. 17·7.—Elements of a linear-array antenna.

**17·6. Dispersion or "Squint."**—The direction of the beam from a linear array varies with frequency. For this reason, an error will be made in determining azimuth by radars that have linear arrays as antennas or antenna feeds, if the frequency of the beacon replies differs from that of radar echoes. This error is sometimes called "squint." A method of computing it is described below.

In Fig. 17·7, *A* and *B* are two successive elements of a linear array, with separation *S*. The arrows indicate the direction of the wave normal. The wavefront *AC* will make the angle $\theta$ with *AB* and the phase at *A* and *C* is the same. Let us assume that *A* and *B* are dipole elements in a waveguide in which the wave travels from $A$ to $B$, and that their phases are opposite.

Let $\lambda$ and $\lambda_g$ be the free-space wavelength and guide wavelength, respectively, and $\lambda_0$ and $\lambda_{g_0}$ be the values of the same quantities at the frequency for which the antenna is designed. Then if $S = \lambda_{g_0}/2$, the phase at $A$ and $B$ is the same, $\theta = 0$, and the wave normal is perpendicular to the array at the design frequency.

At any other frequency, $\lambda \neq \lambda_0$, and the condition for equality of phase at $A$ and $C$ is

$$\frac{S}{\lambda_g} + \frac{S \sin \theta}{\lambda} = \frac{1}{2},$$

from which

$$\sin \theta = \lambda \left( \frac{1}{\lambda_{g_0}} - \frac{1}{\lambda_g} \right).$$

Using the relation

$$\lambda_g = \frac{\lambda}{\left[ 1 - \left( \frac{\lambda}{\lambda_c} \right)^2 \right]^{1/2}},$$

where $\lambda_c$ is the critical guide wavelength, we have also

$$\lambda_{g_0} = \frac{\lambda_0}{\left[ 1 - \left( \frac{\lambda_0}{\lambda_c} \right)^2 \right]^{1/2}},$$

and introducing a new variable $l = \lambda/\lambda_0$ and a constant $l_c = \lambda_0/\lambda_c$, we have

$$\sin \theta = l(1 - l_c^2)^{1/2} - (1 - l^2 l_c^2)^{1/2}.$$

Differentiating with respect to $l$,

$$\cos \theta \frac{d\theta}{dl} = (1 - l_c^2)^{1/2} + \frac{l l_c^2}{(1 - l^2 l_c^2)^{1/2}}.$$

When $\lambda \approx \lambda_0$, $l \approx 1$, $\cos \theta \approx 1$, and the above reduces to

$$\frac{d\theta}{dl} = \frac{1}{(1 - l_c^2)^{1/2}} \text{ near } \theta = 0. \qquad (1)$$

For commonly used rectangular guide, $l_c = \lambda_0/\lambda_c = 0.7$, and $d\theta/dl = 1.4$. For a change $\delta l = 0.01$, this gives $\delta\theta = 0.014$ radian $= 0.8°$. Hence, if beacon and radar signals are presented simultaneously, there will be a relative azimuth error of 0.8° for a 1 per cent difference in frequency. If a coaxial-line array is used instead of waveguide, Eq. (1) reduces to $d\theta/dl = 1$, and the shift is correspondingly smaller.

In electrical scanning waveguide arrays in which the scanning is accomplished by varying the width of the guide, the azimuth error due to off-frequency response varies with the azimuth, being a minimum on one side of the forward direction and a maximum at the other extreme where the guide wavelength is very large. It appears to be impractical to get beacon azimuth information from such a system unless the beacon frequency happens to be very close to that of the radar transmitter.

When space and weight are not of primary consideration, as in heavy ground radar sets, the azimuth error due to off-frequency reply can be eliminated by using a separate beacon receiving antenna which can be aligned independently of the radar antenna.

## INDICATORS

**17·7. Sweep Speed and Code Legibility.**—Range coding (see Sec. 5·12) has been used widely as a means of beacon identification. Several conditions must be fulfilled if this type of code is to be identified readily on a cathode-ray-tube display. First, if several different pulse spacings are employed, they must be sufficiently different so that the observer can distinguish between them quickly. Common practice has been to restrict the number of different spaces to two, the ratio of the longer to the shorter having a value of about 2.5. This makes it possible to identify a code group by means of the *relative* spacing alone, provided that neither codes with long spaces only, nor those with short spaces only, are used. If a greater variety of spaces is used, means must be provided on the radar indicator for measuring the intervals.

The code must be recognizable on the slowest sweep; thus allowing a reasonable safety factor, the smallest code space must be at least twice the minimum spot size when the cathode-ray tube is focused carefully. For example, for a 5-in.-diameter PPI tube, the minimum spot diameter is about 0.016 in. If the shortest code space is 1 nautical mile, then it will be resolved clearly on a sweep of $\frac{1}{2 \times 0.016} = 31$ nautical miles per inch. Long ranges may be measured by means of delayed sweeps, thus avoiding the need for very slow sweeps.

The code cannot be identified unless the whole code is visible on the cathode-ray tube. For example, a fast sweep of 1 mile/in. may be desired for accurate ranging, while a six-pip beacon code may extend over 10 miles. Alternative solutions are to switch to a slow sweep for purposes of identification, or to provide a separate indicator with a slow sweep. The latter is done in the SCR-584, in which tracking is done on a 1-mile sweep J-scope and identification on a 16-mile J-scope or a PPI with 50- or 100-mile sweeps.

**17·8. Signal Intensity.**—The following remarks apply to intensity-modulated indicators and particularly to airborne search radars. A number of factors influence the intensity of beacon signals which are considerably above noise level, without greatly changing the minimum signal strength which can be detected. An intense signal is desirable because it is less tiring to detect and observe it.

It has been found that the optimum video level for beacon reception is frequently higher than that for search operation. In the latter case, a fixed low video level makes it possible to see weak signals close to strong signals, thus giving maximum discrimination in land painting. In beacon reception, since all pips of a beacon code have the same intensity, the i-f gain control can be used to reduce signal strength if "blooming" causes loss of definition.

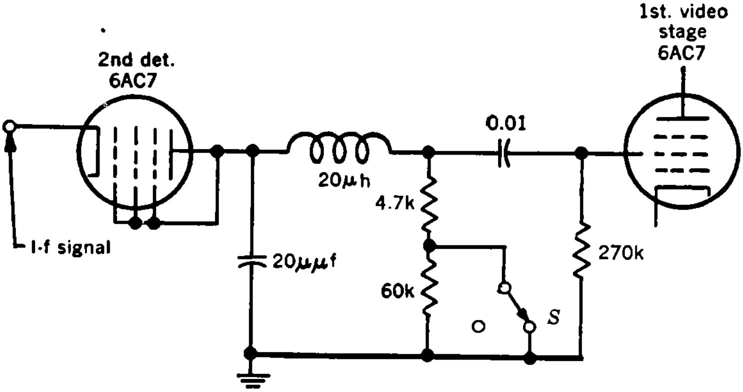


FIG. 17·8.—Video-stretching circuit. The switch $S$, when open, inserts the 60-k resistor which increases the discharge time of the 20-μμf condenser.

Each individual pip of a code is formed on the indicator tube by a number of pulses, occurring on successive sweeps. In order to take maximum advantage of the build-up on a screen that gives a persistent image, these pulses should overlap. That is, the displacement of the spot corresponding to the angular rotation of the scanner between successive sweeps should be less than the diameter of the spot of the cathode-ray beam. For a PPI indicator, this requires

$$\frac{\pi DS}{N} < d$$

where $S$ is the number of scans per second over the beacon, $D$ the useful diameter of the tube face, $N$ is the repetition rate, and $d$ is the diameter of the spot. For a given diameter of tube and repetition rate, then, the signal intensity will be higher if $S$ is made small enough to satisfy this relation.

Another method of increasing signal intensity is known as video stretching, which is discussed in Sec. 2·8. Suppose a range sweep of 40 nautical miles/in. is used on a PPI tube, the minimum spot diameter being 0.016 in. at sharpest focus. If the duration of the pulse from the beacon is 0.5 $\mu$sec, this is equivalent to 0.001 in. on the screen, so the actual length of the signal on the screen is determined almost entirely by the spot size. Figure 17·8 shows a modification made on a standard airborne radar receiver to introduce video stretching. A resistor $R_1$ is added in the plate circuit of the second detector. The slope of the leading edge of an incoming signal is unaltered, but the trailing edge is considerably extended by lengthening the time constant for discharge of condenser $C_1$. This increases the effective length of an incoming 0.5 $\mu$sec pulse by a factor of approximately 5. There is little loss in signal definition, which is still largely determined by the spot size, but the intensity is increased considerably. There is no gain in signal-to-noise ratio because noise is also stretched in the same way.

**17·9. Range Accuracy: Correction for Beacon Delay.**—All beacons return their first signal after a slight delay corresponding to the time required for the pulse to actuate their various circuits. In particular, a beacon which incorporates pulse-width discrimination, accepting pulses only if they are longer than $\tau$ $\mu$sec, must obviously have a minimum delay greater than this. Representative figures for the total delay range from 1 to 5 $\mu$sec. Variations in delay range from $\pm 0.5$ to $\pm 0.1$ $\mu$sec.

Aside from this inherent delay, the accuracy with which a radar can measure ranges to a beacon is determined by the ranging circuit and the indicator display. The delay in the beacon, assuming that it is constant, can be allowed for by incorporating a delay in the starting of the radar range sweep. This is best done by adding an accurate fixed delay in the radar sweep, and using a delay "trimmer" in each beacon to permit adjustment to the standard value. The delay in the range sweep should be connected to the search-beacon switch, so as to be out of the circuit on search operation. However, in systems having dual presentation, in which both radar and beacon signals are displayed on the indicator simultaneously, it is necessary to introduce a delay both in the video channel of the search receiver and in the range sweep, so that both radar echoes and beacon signals will be accurately registered in range. The required delays are small enough to be achieved with delay lines; these must have adequate bandwidth.

On a J-scope indicator with 1-mile circular sweep, used in some accurate ranging radars, a range setting to 0.1 $\mu$sec (50 ft) or better is easily possible. Allowing for a $\pm 0.1$-$\mu$sec delay variation in the beacon, maximum range errors of 100 ft or less are easily attained with precision range circuits. In the case of intensity-modulated presentation, setting errors are slightly greater; the "edge" of a signal is slightly less well defined, even on an expanded scale. Expansion of the scale is worth while up to the point at which the width of the pulse on the indicator is considerably greater than the size of the spot.

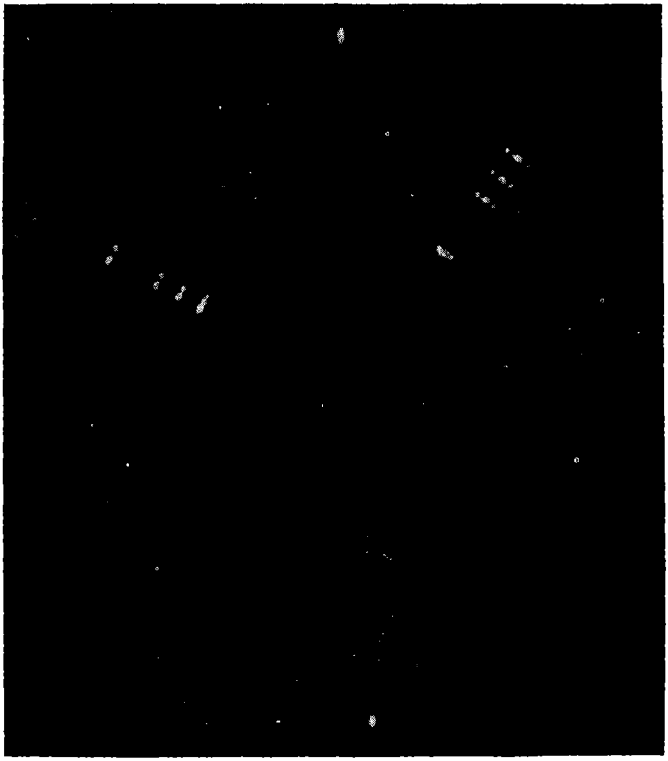


FIG. 17·9.—Split azimuth presentation. The sweep length is 12 miles. The sweep for the azimuth sector containing the beacon whose code is 3-1 is delayed 114 miles. The sweep for the other beacon (code 1-3) is delayed 74 miles. Precise range measurements on both beacons are thus made possible at the same time.

**17·10. Split Azimuth Presentation.**—In an *H*-system of navigation, the distances to two beacons at known locations are measured by a radar in order to obtain a fix. With PPI presentation, range delays must be introduced in order to get an expanded scale. In order to see two beacons at different ranges, so-called "split presentation" has been used. Azimuth switches, the positions of which can be varied, are used on the PPI to switch between independently variable accurate delays in two sectors.

For example, suppose the positions of two beacons are 87 miles at 90° and 133 miles at 210°. The azimuth switches might be set to switch at 150° and 300°, with delays of 85 and 130 miles in the two sectors. Then on a 5-mile delayed sweep, the beacons would appear at 2 and 3 miles, respectively, and ranges could be measured much more accurately than on a long common sweep. Figure 17·9 shows such a split, delayed presentation.

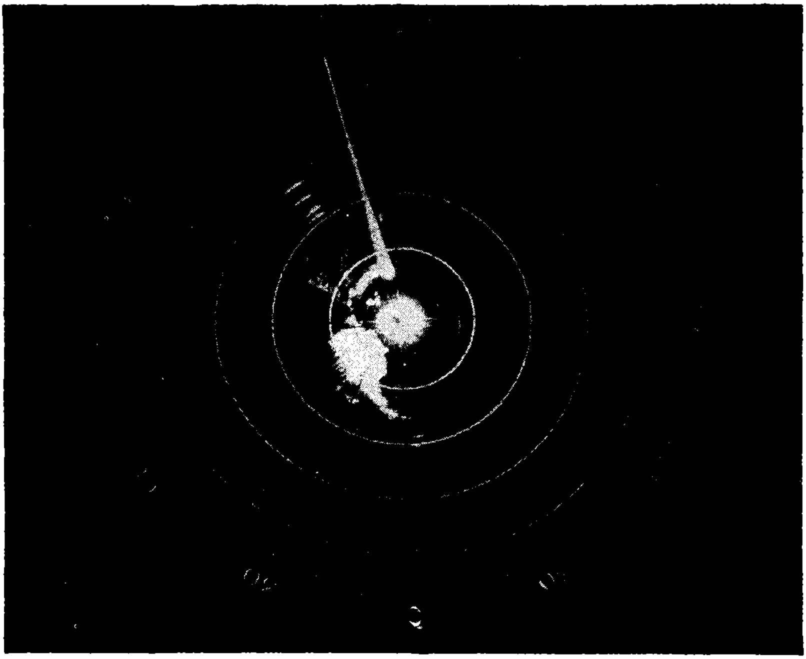


FIG. 17·10.—Simultaneous beacon-radar PPI presentation on AN/APS-10. The beacon (code 2-1-1) at 21 miles, 145°, is the South Weymouth, Mass., 3-cm installation. The radar presentation shows the coast line near Plymouth. Range circles are at 10-mile intervals.

*Independent vs. Dual Presentation.*—In the earlier beacon-radar systems, a function switch gave *either* radar *or* beacon signals on the indicator. It was early realized that in many situations it would be advantageous to have both on the screen simultaneously. Consider, for example, an aircraft navigating cross-country by radar, looking for an airport near a city. The city and its surroundings show up on radar in many cases as a confused mass of signals. If there is a beacon at the airport, the operator can switch to beacon operation and home on it, but he then loses the benefit of radar. With both together on the screen, the beacon signal clearly locates the airport with respect to the radar map of the city. Or again, a plane may be homing on a beacon from a great distance, yet wish to use its radar to navigate around storms or to warn of approaching aircraft or mountain tops. Figure 17·10 illustrates simultaneous beacon-radar reception on an airborne radar set.

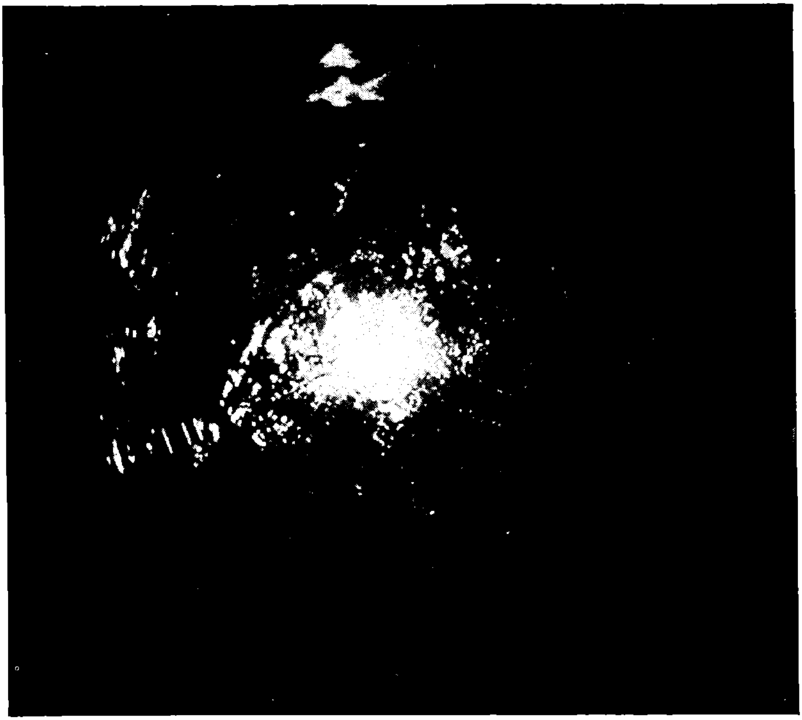


FIG. 17·11.—Simultaneous beacon-radar PPI presentation on a 10-cm ground radar set. An airborne beacon (code 2-1) is seen at 20 miles. Range circles are at 10-mile intervals.

Several methods have been suggested for obtaining dual presentation. If a single receiver is used for both search and beacon, it is practical to switch from beacon to search on alternate scans (as in Fig. 17·10), provided the scanning rate is neither too fast nor too slow. Another suggestion is a much more rapid switch, either on alternate pulses or on alternate groups of pulses. Both systems involve some presentation loss on both radar and beacon.

When weight, space, and complexity permit, the best solution involves separate receivers for beacon and search functions, with mixing after the video amplifier (Sec. 17·3). In this manner, selector-switch choice of search only, beacon only, or search and beacon simultaneously is available. The greatest benefits of this flexibility are realized with multitone presentation. For example, the beacon video-output signal may be limited at a value 50 per cent above that of the search receiver. Beacon signals will then be clearly distinguishable even if superimposed directly on search signals.

FIG. 17·12.—Radar presentation alone (same as in Fig. 17·11).

Figures 17·11, 17·12, and 17·13 illustrate dual presentation of an airborne beacon on a ground radar set. Figure 17·12 shows the ground echo pattern. Figure 17·11 illustrates dual presentation, with the signal from an airborne beacon in among the permanent ground echoes on the screen of a ground radar. Two-tone presentation was not used; the beacon signal is, therefore, not easily picked out. Figure 17·13, a photograph taken shortly after Fig. 17·11, shows the same beacon at a slightly greater range, with the radar echo receiver turned off. All but the closest ground signals have disappeared.

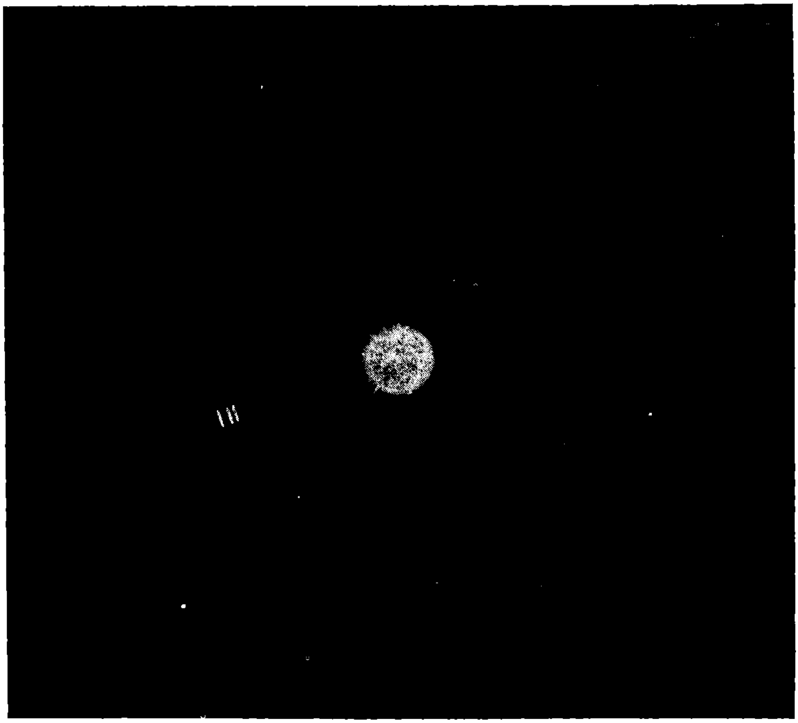


FIG. 17·13.—Beacon presentation alone. This picture, taken slightly later than Fig. 17·11, shows the beacon further away.

# CHAPTER 18

# THE DESIGN OF INTERROGATOR-RESPONSORS

BY G. O. HALL AND M. D. O'DAY, WITH CONTRIBUTIONS BY J. R. LIEN

All interrogator-responsors include antenna systems, drivers, modulators, transmitters, and receivers, the design of which is determined largely by the specific application intended for the device. In some cases it is necessary to add other elements, such as coordination circuits and indicators. Figure 18·1 shows a block diagram of a typical interrogator-responsor; only those components common to all units of this type are indicated. The similarity to a radar is evident. If the transmitting and receiving frequencies are the same, and if ranging circuits and an indicator are included, the interrogator-responsor is a radar.

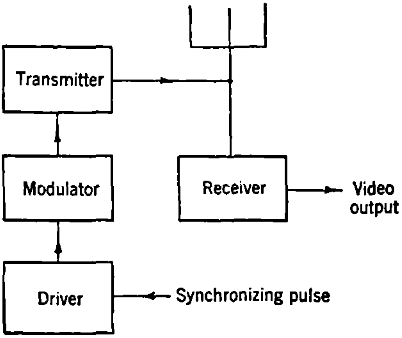


FIG. 18·1.—Block diagram of a typical interrogator-responsor.

The various circumstances under which radar sets are used as interrogators and under which interrogator-responsors are used alone or in conjunction with radar sets have been discussed in Part I.

Several practical considerations often make an interrogator-responsor quite different from a radar. The element of expense may be important; interrogator-responsors will often be less expensive than radar sets. Because one-way transmission is involved, the power of the interrogator can be less than that of a radar transmitter. For the same reason, the transmitter of the beacon can be designed so that the responsor can be rather simple.

## INTERROGATOR-RESPONSOR USED ALONE FOR RANGE-ONLY INFORMATION

Since applications of operation of this type usually demand precise ranging, reduction of system delays and methods for accurate measurement of range are prime design considerations. Accuracies greater than ±20 yd (±0.12 μsec) are required in many cases. Unless both the interrogator-responsor unit and the beacon are integrated in design to reduce over-all errors to a minimum, such accuracy is impossible.

**18·1. Antenna.**—Because no azimuth information is desired, the antenna can be stationary with an omnidirectional azimuth pattern. The improvement in signal-to-noise ratio obtainable with a directional array should be considered, however, as many applications will permit some directivity of the antenna without sacrificing the general flexibility of the system. Even if an omnidirectional azimuth pattern is needed, appreciable gain in the vertical radiation pattern may still be desirable. For applications such as the establishment of an accurate baseline in surveying, the direction of the interrogated beacon is known at least approximately and the azimuth pattern can be narrowed to give improved performance. The choice between a single antenna or separate antennas for transmitting and receiving depends upon specific design characteristics—in particular the power, the frequency, and the relative advantages of duplexing and using separate antennas.

**18·2. The Interrogator.** *Modulator.*—The modulator for the interrogator-responsor must supply output pulses with steep leading edges. A rapid rise of the modulator output pulse is essential not only for reducing fixed system delay but also for reducing the variation in delay introduced by replacing oscillator tubes. Because a minimum potential must be applied to an oscillator tube before it will oscillate, a slow rise of the pulse will delay the start of oscillations. A slight difference in the characteristics of replacement tubes will alter the threshold potential and cause variation in delay if the applied voltage pulse rises slowly.

Plate- or cathode-pulsing of triode oscillator tubes is preferable to grid pulsing because grid pulsing introduces greater delays and variation of delays with change of tubes.

*Transmitter.*—Frequency stability and power output are important. Delay also depends on frequency stability. The beacon receiving the interrogations has a certain receiver pass band; some threshold value of voltage delivered by the receiver is required before the transmitter will be triggered. Should the interrogating r-f pulse be off frequency, the beacon receiver will have lower sensitivity and the r-f interrogation will have to reach a higher amplitude before sufficient receiver output is obtained. Inasmuch as every r-f interrogation and every video signal from a receiver will have a sloping leading edge, frequency shift will result in increased delay in the reply.

Furthermore, the signal level of the system should be high so that r-f pulses of large amplitude will reach the beacon receiver. If the signal is well above the level necessary for triggering the beacon, delay variations will be at a minimum. High signal level will also decrease the variation in delay resulting from shift of frequency.

*Synchronization.*—Synchronization and methods for measuring range accurately must receive careful consideration in design. A crystal oscillator is the usual basic timing device used for high precision; operational accuracies of ±50 ft are obtained. With careful design and calibration, the errors can be reduced to less than ±10 ft.

Two types of crystal-oscillator-controlled ranging circuits are used with interrogator-responsors.[1] In one type the frequency of the crystal oscillator is divided by means of blocking-oscillator dividers to provide accurately spaced timing pulses that are used to synchronize the transmitter and indicator circuits. The time interval between pulses is too great for direct use for accurate range measurement; calibrated vernier delay circuits such as phantastrons or delay multivibrators are utilized to interpolate between timing pulses.

The other system utilizes sine-wave frequency division and calibrated phase shifters to vary the time relation of selected cycles of the crystal-oscillator output. The phase-shifting type of circuit is generally more accurate but is more complex and heavier. Selection of the synchronizing and ranging circuit to be employed will depend upon the accuracy demanded of the system and the limitations on space and weight.

**18·3. The Responsor.** *Receiver.*—A receiver with constant delay and minimum distortion is a necessity for accurate range measurement. The distortion and variation in delay inherent in a superregenerative receiver preclude its use for this application. Tuned radio frequency and superheterodyne receivers have suitable characteristics if properly designed. In order to reproduce the envelope of the received pulse, adequate i-f and video bandwidths must be provided. The frequency of operation will be an important factor in the choice of receiver type.

*Indicator.*—The indication should be such as to make accurate ranging easy. Since no azimuth information is required, deflection-modulated presentations such as type A or J are employed. In order to obtain an accurate range measurement, the final determination of range is performed on an expanded presentation. For high accuracy the expanded sweep speed should be about 1 in. per 1000 ft.

It is necessary that this expanded sweep cover the period of time during which the beacon reply is received. This can be accomplished either by advancing the time of the transmitted pulse or by delaying the start of the sweep with respect to the sine wave from the crystal oscillator. For example, if a beacon reply is being received at a range of 101 miles and the expanded sweep is 2 miles long, the reply will be visible on the scope if the transmitted pulse is advanced a time equivalent to 100 miles or if the sweep start is delayed by 100 miles.

---

[1] For details of these systems see *Electronic Instruments*, Vol. 21, Secs. 6·4, 6·5, 6·9, 7·4, and 7·11 through 7·16, Radiation Laboratory Series.

If the transmission of intelligence along with the beacon interrogations is required, the modulation circuits can be modified by the methods outlined in Chap. 11.

For some applications, it may be desirable to have a common location for both an interrogator-responsor and a beacon. In this event, the units can be consolidated to save space, weight, and cost.

*Other Display Methods.*—A simplification of the equipment is possible by eliminating the cathode-ray tube and using a lamp, meter, or "magic-eye" tube. Such circuits permit the display of only one beacon response at a time.

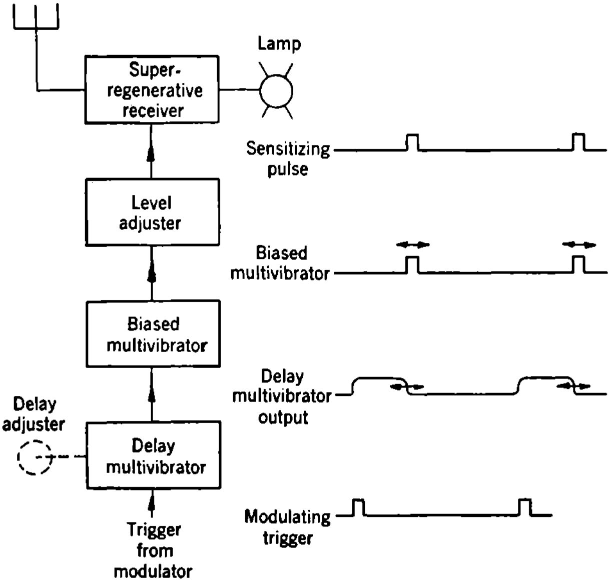


FIG. 18·2.—A simple superregenerative responsor. A lamp display is used; the lamp lights only when the response appears at a range included within the short sensitizing gate.

A simple responsor using a lamp to locate a beacon in range is shown in Fig. 18·2. The superregenerative receiver is biased to a point at which it is insensitive to incoming replies. A sensitizing pulse overcomes this bias and thus makes the receiver sensitive for the duration of the pulse. If a reply is received during the sensitizing pulse, the superregenerative tube will oscillate and plate current will flow through the lamp in its cathode. Each time an interrogating pulse is transmitted, the synchronizing pulse triggers a delay multivibrator, which in turn triggers a delayed sensitizing pulse. The length of the delay is determined by the position of a calibrated range dial. A switch is employed to change the constants of the multivibrator so that there is produced a long sensitizing pulse for searching (this waveform is not shown in the diagram) for beacons. This pulse extends from the end of the synchronizing pulse for a time sufficient to sensitize the receiver for the maximum operating range of the system. In order to secure proper sensitivity for the receiver, the output pulses from the delay multivibrator are adjusted in amplitude by the level-adjusting circuit, which can be a clipping diode or a voltage-dividing potentiometer.

Operation of the equipment is rather simple. With the receiver tuned to the proper frequency and the switch on "range," the sensitivity control is adjusted until the lamp is on the threshold of operation on noise. The range dial is oscillated back and forth to make certain that this adjustment is not being made while a reply pulse is present in coincidence with the sensitizing pulse. The switch is then thrown to "search" to produce the long sensitizing pulse and the antenna is rotated until the lamp lights and thereby indicates that a beacon has been received. The switch is then thrown to "range" and the range dial is rotated until the lamp lights. Range is read directly from the calibrated range dial.

**18·4. Combined Interrogator-responsor-beacon.**—In lightweight range-only systems, it is possible and economical to design the system so that the same equipment can be used as either an interrogator-responsor or as a beacon simply by throwing a switch. The equality of transmitter powers and receiver sensitivity thereby attained assures a balanced system. Lightweight systems of the kind, weighing less than 25 lb, have been built for ground-to-ground use in surveying.

## INTERROGATOR-RESPONSOR USED ALONE FOR RANGE AND AZIMUTH INFORMATION

Present applications for operation of this type do not demand accuracies in ranging comparable to the accuracies demanded of interrogator-responsors for range-only information. The less rigorous requirements allow simplification of the design with resulting economy.

**18·5. Antenna.**—The direction of a beacon being interrogated can be determined by the use of a single directional antenna or by lobe switching. The choice of the antenna system will depend upon frequency, allowable weight and size, azimuth accuracy demanded, and any special design characteristics imposed by the application. In general, lobe switching will give greater azimuth accuracy than a single directive antenna of the same beamwidth.

For ground or mobile installations, the antenna must be rotatable, with suitable azimuth-indicating equipment to show the direction in which the array is pointing. On aircraft, the antennas may be fixed and the aircraft rotated to alter the directions in which the array is pointing. Azimuth information can then be obtained from the aircraft compass, and special azimuth-indicating equipment will not be needed. A fixed antenna array pointing dead ahead reduces the weight, complexity, and cost of the system.

If the frequency is low, an array producing a narrow beam will have large physical dimensions; such an array is permissible for most ground installations but impractical for airborne equipment. When the frequency is high, a suitable directive antenna for aircraft is generally of moderate size and the use of lobe switching or two antennas is not required unless extreme accuracy of azimuth is required.

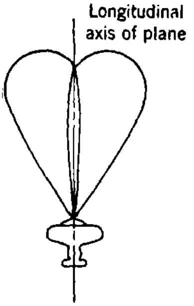


FIG. 18·3.—Direction-finding by lobe-switching in an airborne interrogator. The two overlapping antenna radiation patterns have equal intensities only in one direction, usually the longitudinal axis of the plane.

*Lobe-switching Systems.*—One airborne system, operating at the comparatively low frequency of 200 Mc/sec, avoids the large size of a single highly directional antenna by using two small directional antennas. The two antennas are attached rigidly to the aircraft and so oriented that the antenna radiation patterns overlap with the line of equal signal strength coincident with the longitudinal axis of the plane, as shown in Fig. 18·3. Transmission and reception are alternated between the two antennas at about 15 cps by means of a motor-driven switching unit. Azimuth accuracy is about ±3° for 45° half-power widths.

Airborne lobe-switching systems with fixed antennas suffer from one serious defect, however. In the presence of cross wind, the ground track of the aircraft is not along the direction of the aircraft heading. In homing, the object of the aircraft navigator is, of course, to find a heading such that the ground track of the aircraft will pass through the beacon. If the heading of the aircraft differs from the ground track, which is the case when any component of the wind exists at right angles to the ground track, the two pips obtained by lobe switching will not be equal in amplitude when the aircraft has the correct ground track.

Figure 18·4*a* shows the appearance of the L-scope when the aircraft has the heading that will make its ground track pass through the beacon. Figure 18·4*b* shows what happens if the navigator flies a course such that the two deflections are equal. If the plane is continually oriented to keep the deflections equal, the course flown will be a spiral of continuously decreasing radius of curvature. Homing under these circumstances is difficult, and sometimes impossible.

The difficulty can be cured either by the use of suitable procedures of navigation or by the use of a movable antenna system, as shown in Fig. 18·4*c*. Suitable navigational procedure with fixed antennas requires the navigator to find and keep a heading at which the deflections are unequal, but their ratio remains constant, as shown in Fig. 18·4*a*. This is difficult in practice. A movable-antenna system permits the use of equal deflections to give the proper heading, as in Fig. 18·4*c*. In this case, the navigator rotates the movable array through an angle equal to the drift angle. The proper course is then indicated by equal deflections on the scope.

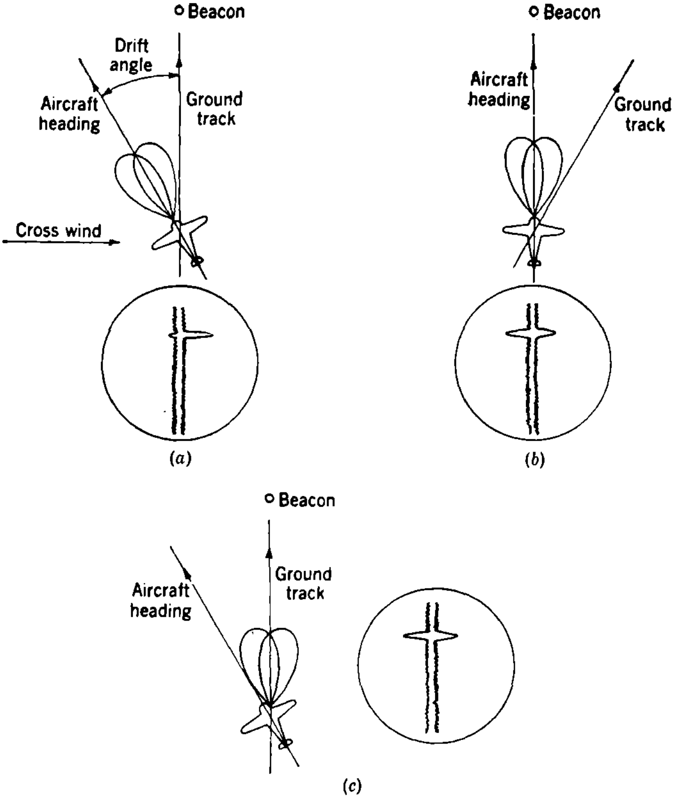


FIG. 18·4.—The effect of cross wind on homing with airborne lobe-switched antennas. In (*a*) the unequal deflections, resulting when the aircraft heading differs from the ground track, are shown. In (*b*) is shown the ground track resulting from an attempt to home by matching pips. A cure for this difficulty, the installation of movable lobe-switched antennas, gives the result shown in (*c*), where the antennas are displaced through the drift angle. In case (*c*) alone is homing by pip matching possible.

Although movable antennas are more complicated and more difficult to install than fixed antennas, satisfactory homing under conditions of strong cross wind is almost impossible without them. In addition, movable antennas make it possible to obtain fixes in range and azimuth on other beacons off course. The best solution available is, of course, the use of a rotating microwave antenna which gives a map display.

**18·6. Interrogator.** *Modulator.*—Because only moderate range accuracy is required, the shape of the r-f interrogating pulse is not critical. Any one of a wide variety of modulation methods may be employed. Inasmuch as a constant pulse-repetition frequency is not essential, a simple and economical method for modulation is self-pulsing of the oscillator. Figure 18·5 shows the basic elements of a self-pulsed oscillator[1] with accompanying grid and cathode waveforms.

Each time the tube oscillates, the plate current produces a voltage drop across the cathode resistor $R_k$, and synchronizing pulses are obtained as shown by the cathode voltage waveform. The condenser $C_k$ is of small size so that it will bypass the r-f. The r-f pulses produced by a self-pulsed oscillator are more or less rounded in shape, but provide leading edges accurate enough for most applications. The r-f components are omitted from Fig. 18·5 because any triode oscillator can be self-pulsed in this way.

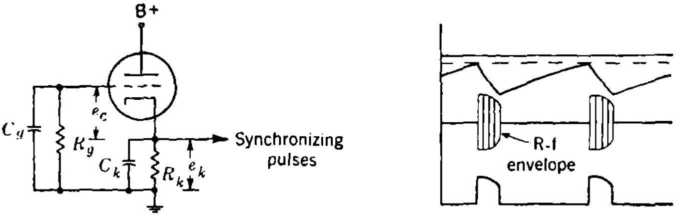


FIG. 18·5.—Self-pulsed oscillator. $R_gC_g$ determines the repetition period. The pulses are not square, but decrease in amplitude.

More power output can be obtained from the triode oscillator if it is plate-pulsed; most triodes will operate with about 50 per cent higher plate voltage when plate-pulsed than when grid-pulsed. The addition of another tube operating as a free-running blocking oscillator with pulse-transformer output can be used to provide plate-modulation pulses for the oscillator, as shown in Fig. 18·6. The pulse transformer in the plate of the blocking oscillator can often have a stepup ratio and thus reduce the high-voltage requirement. The grid condenser and resistor, $C_g$ and $R_g$, determine the pulse-repetition frequency and pulse width for the system in the same way as for the self-pulsed oscillator. The require-

---

[1] Such an oscillator is sometimes called a "squegging" oscillator.

ment that the blocking oscillator tube handle both the power necessary for the triode oscillator and the power for supplying its own losses must be taken into consideration in selecting a tube for this purpose. With a good pulse transformer, the shape of the r-f output pulses will be better than those delivered by a self-pulsed oscillator. Synchronizing pulses for the indicator equipment can be obtained from a resistor placed either in the blocking oscillator or in the r-f oscillator cathode.

*Transmitter.*—The r-f oscillator must be stable in frequency and provide adequate power output. The required range of operation and characteristics of the beacons to be interrogated will determine the degree of frequency stability required and the power output needed. The pass band of the beacon receiver will determine the frequency deviation allowable for a given power in the interrogating pulses. The power output from the interrogator should be sufficient to trigger a beacon consistently at the maximum range required for the system.

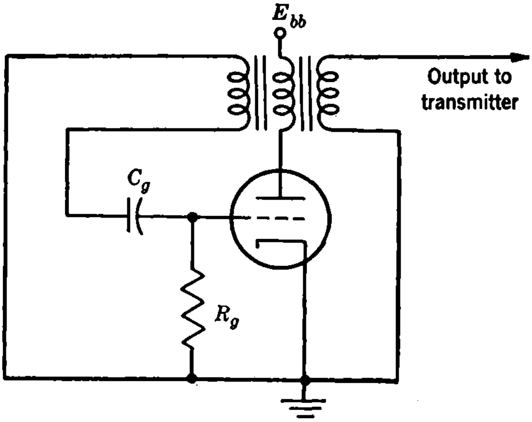


FIG. 18·6.—Simple blocking-oscillator modulator.

**18·7. Responsor.** *Receiver.*—The receiver may be of any type suitable for reception of pulses. Its sensitivity will be dictated by the signal voltage obtained from a beacon at the maximum range of the system. When high range accuracy is not required, the receiver may be very simple.

The superregenerative receiver is a simple, low-cost receiver suitable for many applications. Gains of the order of $10^6$ can be obtained by a single stage; its comparatively small size and light weight make it particularly desirable for airborne installations.

*Indicator.*—Either an A-scope or an L-scope, depending upon the method for obtaining directivity with the antennas, is suitable for displaying the beacon replies. For a low-cost system, the expense of a cathode-ray tube and its associated circuits can be avoided by the use of a range-gated indicating device, such as a bell, lamp, or meter (as described in Sec. 18·3) with, however, the corresponding sacrifice of the ability to see more than one beacon at a time.

Some applications for interrogator-responsors may demand that communications be transmitted along with or by the interrogating pulses. This is discussed in Chap. 11.

## RANGE-ONLY INTERROGATOR-RESPONSORS USED WITH RADARS

It is often desirable to display beacon replies and radar echoes on the same indicator in order to associate a specific beacon reply with the radar echo from a particular ship or plane. Since a beacon system replies on a fixed frequency and radars may operate over a wide range of frequencies, an interrogator-responsor may be employed to challenge the beacons if the system is to be used with many different radars.

To simplify the interrogator-responsor, an omnidirectional antenna may be utilized. Because the correlation between beacon replies and radar echoes is by range only when an omnidirectional antenna is used, it is essential that there be sufficient difference in range of radar echoes to resolve any ambiguity. For example, three radar echoes and a beacon reply might appear at the same range and it would be impossible to decide which of the echoes should be associated with the beacon. If the echoes represent three airplanes, however, their ranges will change in a comparatively short time; the beacon-carrying plane can then be identified.

**18·8. Synchronization.**—All interrogator-responsors that work with radars must have circuits that coordinate the information coming from the radar to the interrogator-responsor, and vice-versa. The first function of such a circuit is to accept synchronizing pulses from the radar for triggering the signals from the interrogator. The pulse-repetition frequency of the interrogator should be held under 500 cps to minimize overinterrogation of the beacons. Consequently, the synchronizing network incorporates circuits to prevent the interrogation rate from exceeding 500 cps, while at the same time it synchronizes the interrogations with individual radar pulses. This is accomplished by circuits that prevent the interrogator from accepting pulses from the radar for a given dead time after it has accepted a synchronizing pulse. Blocking oscillators, step chargers, and multivibrators have been used for this purpose.

As an example, it is supposed that a blocking oscillator that cannot be retriggered for 2400 $\mu$sec after it has once fired is employed. It is assumed also that this system is used with a radar set that has a repetition rate of 2000 pulses per second. The synchronizing pulses from this radar occur at intervals of 500 μsec. Once the blocking oscillator has fired, the next four pulses occur during the dead time but the fifth pulse (at 2500 μsec) arrives after the dead time and will trigger the blocking oscillator again. Thus, the interrogator will send out pulses synchronized with every fifth pulse from the radar, that is, 400 pulses per second. In this system the pulse-repetition frequency of the interrogator-responsor will vary with the pulse-repetition frequency of the radar. If the radar has a pulse-repetition frequency less than the reciprocal of the dead time, (416.7 pulses per second in our example), the radar and interrogator-responsor will transmit at the same pulse-repetition frequency. If the radar pulse-repetition frequency exceeds the reciprocal of the dead time, the circuit will divide the pulse-repetition frequency of the radar by some integer. At a radar pulse-repetition frequency of 500 cps, for example, it will divide by two.

If it is necessary for the beacon reply to occur at exactly the same range as the radar echo from the beacon-carrying plane or ship, the coordination circuit must compensate for the delays inherent in modulators, synchronizing circuits, and beacons. If the synchronizing pulse is simultaneous with the radar output pulse, there will be a delay in the synchronizing and interrogator circuits of 0.5 to 4 μsec, depending upon the circuits involved. This delay, plus the delay in the beacon, must be taken into account to achieve range coincidence between echoes and beacon replies. It is often possible to obtain a prepulse from the radar synchronizer and thus to compensate for the delay by giving the interrogator-responsor system extra time. If a prepulse is not available, the beacon replies can be displayed on a delayed sweep to obtain range coincidence, or the radar display can be delayed by the same amount. The use of a delayed sweep requires that the radar and responsor share time on the radar indicator. It can be seen that the use of a prepulse or a delayed radar display is simpler and more satisfactory and should be adopted whenever possible. In some cases it may be desirable to have a definite delay between radar echoes and beacon replies. This delay can be incorporated in the coordination circuits.

If the interrogator-responsor is located near a beacon of the same type that it challenges, it may be required to supply a suppression pulse to this beacon to prevent it from replying.

**18·9. Interrogator.** *Transmitter.*—It is usually desirable to have the range of the interrogator-responsor-beacon system greater than the radar range so that beacon replies will be received from beacon-carrying targets at greater ranges than those at which they are detected by the radar. The interrogator power and responsor sensitivity, therefore, must be determined on the basis of maximum range requirements. The required interrogator power and responsor sensitivity may be calculated from the considerations of Chap. 2.

*Modulator.*—If the pulse-repetition frequency of the interrogator-responsor is fairly constant, a simple gas-filled tube modulator is very satisfactory. Modulation that makes plate pulsing possible is preferable. For frequencies below 500 Mc/sec, it usually is possible to attain an oscillator plate efficiency greater than 20 per cent. The pulse power of the modulator can usually be calculated on this basis.

**18·10. Responsor.** *Receiver.*—Both superheterodyne and superregenerative receivers have been used for this application. A well-designed superheterodyne receiver with adequate over-all bandwidth will reproduce faithfully the pulses received from beacons. The superregenerative receiver, however, has advantages of light weight and lower cost which may, in some cases, outweigh its inferior performance. It has been used extensively in airborne interrogator-responsors.

If the output of the responsor is connected to the radar indicator at all times, the responsor must be made inoperative except for the sweep period after each interrogation so that interference from it will not be displayed continuously on the radar screen. If a superregenerative receiver is used, it must be suppressed in this way even if it is not continuously connected to the radar indicator. Otherwise, radiation from the receiver can enter the radar and might produce interference.

*Indicator.*—The use of the radar indicator for combined display of beacon replies and radar echoes assists materially in reducing weight and space requirements in aircraft. If the display is on an A-scope, the beacon replies can be introduced into the indicator with polarity opposite to the radar echoes. This will cause the deflections from beacon replies to be opposite to those for radar echoes and thus make them easy to distinguish.

Because the pulse-repetition frequency of the interrogator is usually a fraction of that of the radar, it is possible to share time on the indicator between radar echoes and beacon replies. This reduces noise introduced from the responsor since the responsor output is blanked while radar information is presented.

## INTERROGATOR-RESPONSOR USED WITH RADARS FOR RANGE AND AZIMUTH INFORMATION

The azimuth of beacon replies can be determined either by the use of a directional antenna on the interrogator-responsor or by the employment of beacons that reply only when interrogated simultaneously by the radar and the interrogator-responsor. In the latter case, the directivity of the radar interrogation is used.

**18·11. Design Considerations.** *Antennas.*—If a directional antenna is used for fixed ground, mobile, or ship installations, the antenna is rotated in synchronism with the radar antenna so that both antennas point in the same direction. The simplest method for obtaining this synchronism is to mount the interrogator-responsor antenna with the radar antenna so that both rotate together. Because aircraft can be oriented easily in azimuth, it is common practice in aircraft installation to use fixed antennas that interrogate a sector directly ahead of the aircraft. If beacons requiring coincident interrogation are employed, the interrogator-responsor antenna can be omnidirectional; the improvement in signal-to-noise ratio obtainable with a directional interrogator-responsor antenna, however, makes its use desirable. In this case, the antenna need not produce a narrow beam and a comparatively simple array may be employed.

*Synchronization.*—The problems of coordination between the radar and the interrogator-responsor are identical with those covered in Sec. 18·8, and similar techniques are employed to prevent overinterrogation, range errors, and so on. When coincidence beacons are employed, the characteristics of the beacons dictate the time relation between the radar pulse and the interrogating pulse because the beacons cannot reply unless this time relation is correct.

If azimuth information is obtained by a directional interrogator-responsor antenna, it is often desirable to incorporate circuits that can be used to operate the interrogator-responsor self-pulsed so that the system can be used independent of the radar by providing it with a separate indicator. Operation of this type is discussed in Sec. 18·6.

*Modulator.*—The modulator must be capable of delivering ample power to the transmitter for the requisite r-f power output. A steep rise in the modulator pulse is essential for obtaining accurate timing for use with coincidence beacons. Pulse width will be determined by the system characteristics.

*Transmitter.*—The power output will be determined by the range of the radar, the directivity of the antenna, and the characteristics of the beacons employed. An adequate power margin should be available at the maximum operating range to insure positive triggering.

*Receiver.*—A well-designed superheterodyne receiver gives the best performance. For airborne interrogator-responsors in which weight and size are important considerations, superregenerative receivers have been utilized, but only at the expense of loss in performance.

*Indicator.*—In displaying the beacon replies and radar echoes on the same screen, the problems are the same as those covered in Sec. 18·10 and are solved by the same techniques.

If the system is designed so that it can also be used alone with its own indicator, provision must be made for synchronizing pulses to trigger the sweep-generating circuits.

If communication is required, the methods discussed in Chap. 11 can be employed.

# CHAPTER 19

# TYPICAL BEACON SYSTEMS

By G. C. Danielson, P. A. de Paolo, and F. P. Zaffarano

In this chapter several beacon systems of various types are described briefly. Most of the examples are of existing systems; where no systems of the type desired exist, hypothetical examples have been given. The systems described will, we hope, illustrate the principles and practices of beacon design as presented in the preceding chapters.

## AIRBORNE-RADAR-GROUND-BEACON SYSTEM FOR AIR NAVIGATION

By G. C. Danielson and F. P. Zaffarano

A typical airborne-radar–ground-beacon system is the AN/APS-10 radar in the aircraft operating with the AN/CPN-6 beacon on the ground. The system is designed for navigation of aircraft and provides data on range and azimuth to the aircraft navigator. Navigation can be accomplished either by radar pilotage, using the PPI presentation of radar echoes from the ground, or by taking fixes, using responses from fixed beacons at known locations on the ground.

**19·1. Airborne Radar (AN/APS-10).**—AN/APS-10 is a lightweight, low-power navigational set designed for compactness, simplicity, and reliability, rather than for long radar range. A photograph is shown in Fig. 19·1, and a complete description is given in *Radar Aids to Navigation*, Vol. 2, Chap. 6, Radiation Laboratory Series. The important characteristics of AN/APS-10 as a component of the complete radar-beacon system are listed below.

*General*

Wavelength: 3 cm
Weight (installed): 150 lb, plus cables
Number of operators: one
Power supply: 340 watts a-c, 115 volts, 400 to 1600 cps; 80 watts d-c, 27.5 volts
Packaging: five units—antenna assembly, transmitter-receiver, indicator, synchronizer, and synchronizer power supply

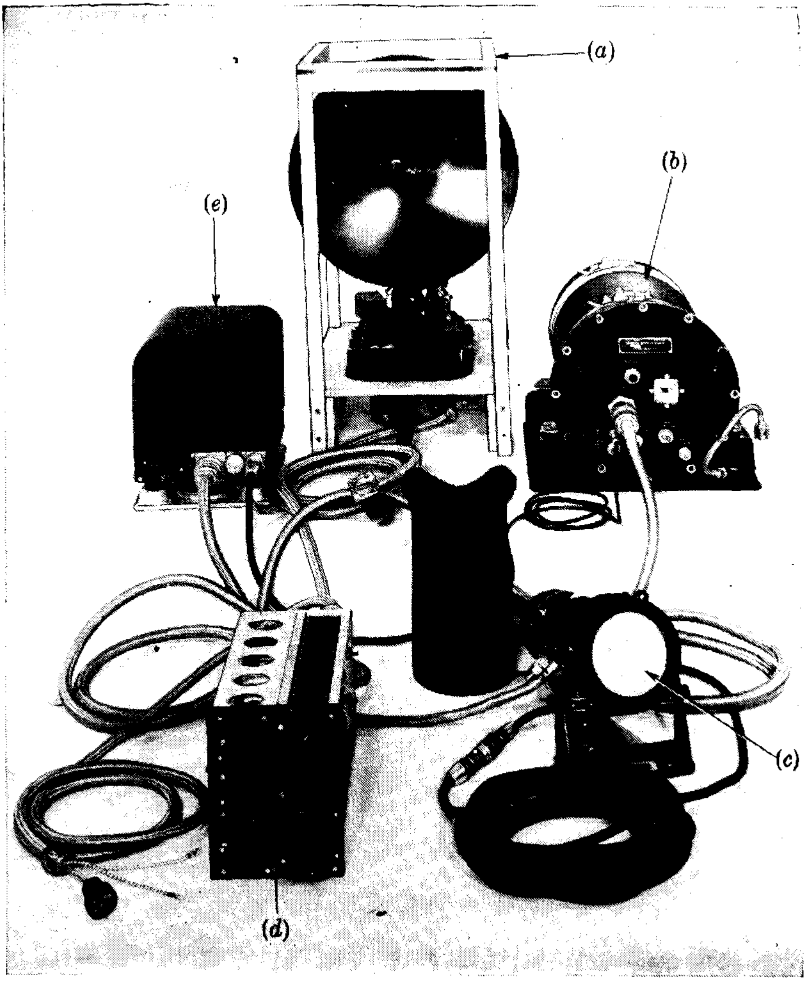


FIG. 19·1.—Assembly of AN/APS-10, lightweight 3-cm radar components. (*a*) Antenna. (*b*) Transmitter. (*c*) Indicator. (*d*) Synchronizer. (*e*) Synchronizer power supply.

*Antenna Assembly*

Type: dipole in 18-in. paraboloid modified for $\csc^2 \theta$ coverage to 7500 ft altitude

Azimuth coverage: circular scan 360°, 30 rpm, half-power beamwidth 5°

Elevation coverage: tilt adjustment $-21°$ to $+3°$

Polarization: horizontal
Gain: 700

*Transmitter*

R-f source: 2J42 magnetron
R-f scatter band (including drift): 9375 ± 45 Mc/sec
R-f pulse power: 10 kw
Pulser: line type; 3C45 thyratron switch-tube
Pulse-repetition rate, radar: 810 pps
Pulse-repetition rate, beacon: 405 pps
Pulse duration, radar: 0.8 $\mu$sec
Pulse duration, beacon: 2.2 $\mu$sec

*Receiver*

Type: superheterodyne
Stages: six i-f, two video
I-f pass band: 5.5 Mc/sec
Sensitivity (minimum discernible signal): 131 dbw on search, 125 dbw on beacon
Special beacon features: separate beacon local oscillator with AFC, slug-tuned TR cavity, broadband ATR cavity, video stretching

*Indication*

Type: 5-in. PPI
Sweeps: 4 to 25 nautical miles, continuously variable on search or beacon;
  50 nautical miles, search or beacon;
  90 nautical miles, beacon only;
  70- to 160-nautical-mile delayed sweep; beacon only
Range marks: 2-mile range marks on 4- to 14-mile sweep;
  10-mile marks on 15- to 25-mile sweep and on 50-mile sweep;
  20-mile marks on 90-mile sweep and on 70- to 160-mile sweep
Azimuth marks: 10° angle marks

**19·2. Ground Beacon (AN/CPN-6).**—AN/CPN-6 is a high-power ground beacon which accepts 3-cm interrogation pulses and transmits a 3-cm range-coded reply. Radar-equipped aircraft are thus provided with range, direction, and identification of the beacon station. A block diagram is shown in Fig. 19·2, and a photograph is shown in Chap. 15, Fig. 15·11.

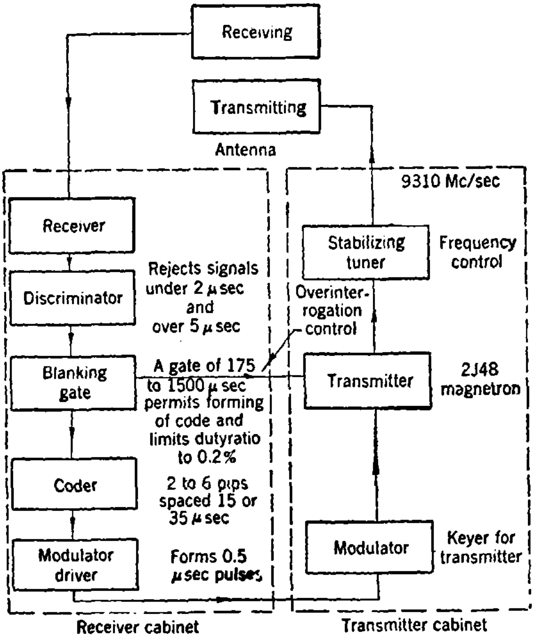


Fig. 19·2.—Operational block diagram of AN/CPN-6 beacon.

The important characteristics of this beacon are listed below.

*General*

Wavelength: 3 cm
Weight (installed): 1300 lb
Number of operators: none
Power supply: 2 kva 115/230 volts, 50 to 70 cps
Packaging: three major units—receiver cabinet, transmitter cabinet, antenna assembly; dimensions of each cabinet, 27 by 18 by 43 in. high; of antenna 5 by 15 by 60 in. high

*Antenna Assembly*

Type: separate receiving and transmitting antennas, each a linear array of 12 slotted-waveguide elements
Azimuth coverage: 360°
Elevation coverage: half-power beamwidth of 5°, with direction of maximum gain 1° above horizontal
Polarization: horizontal
Gain: 20

*Transmitter*

R-f source: 2J48 magnetron
Response frequency: 9310 Mc/sec
R-f stability: ±3 Mc/sec
R-f pulse power: 20 to 40 kw
Maximum duty ratio: 0.2 per cent
Pulser type: 5D21 constant current
Pulse duration: 0.5 μsec
Blanking-gate duration: 175 μsec
Overinterrogation protection: blanking gate increases from 175 to 1500 μsec at high interrogation rates

*Receiver*

Type: superheterodyne with square-wave-modulated local oscillator
Stages: eight i-f, two video
I-f pass band: 12 to 47 Mc/sec
R-f coverage: 110 Mc/sec (between 9320 and 9430 Mc/sec)
Sensitivity (minimum interrogation signal): 97 dbw

*Coding*

Interrogation: pulse-width discrimination; 2- to 5-μsec pulses are accepted
Reply: range-coded reply of two to six pulses with long (35 μsec) or short (15 μsec) spaces; total number of possible codes, 62

*Special Features*

Power-supply control: overload and time-delay relays; automatic reclosing (Sec. 15·8)
Beacon delay: without 5-μsec discrimination, delay modification allows establishment of fixed over-all delay at 4.3 ± 0.1 μsec
Automatic frequency control: AFC modification allows transmitter frequency stability of ±0.5 Mc/sec
Test equipment: complete test equipment supplied including signal generator and synchroscope

**19·3. System Performance.**—The radar performance of AN/APS-10 is described in detail in *Radar Aids to Navigation*, Vol. 2, Chap. 6, Radiation Laboratory Series. Briefly, maximum range for land "painting" is 25 to 30 nautical miles; maximum range for cities is 30 to 50 nautical miles; radar resolution is adequate to distinguish coast lines, lakes, and rivers. The beacon performance of AN/APS-10 operating with AN/CPN-6 beacons is discussed below.

*Coverage.*—Range coverage is limited only by the horizon or by interference nulls. Since the cosecant-squared pattern of the airborne antenna tends to equalize the received-signal strength, the radar can receive beacons at different ranges on a single scan without adjustment of the tilt or gain controls, as shown in Chap. 1, Fig. 1·14. Azimuth coverage is uniform (within 3 db) in all directions from the beacon station. Elevation coverage is uniform (within 3 db) only in the region from $-1.5°$ to $+3.5°$, measured from the beacon station. High-flying aircraft will thus lose beacon signals when directly over, or nearly over, the beacon station.

*Range.*—The reliable beacon range is 160 nautical miles; it is limited by the radar sweep length if not by the horizon. Theoretical reliable range calculations computed as in the example in Sec. 3·11, indicate (1) that the response link is much stronger (18 db) than the interrogation link, and (2) that the interrogation range is about 350 nautical miles. At an actual range of 160 nautical miles, approximately 7 db are thus available for overcoming possible interference nulls. Except in the improbable case of the aircraft flying in a circular course around the beacon, loss of beacon signals from interference nulls will not in general interfere with ordinary navigation.

Minimum range is 400 yd for aircraft flying at altitudes sufficiently low so that beacon reception is not limited by poor overhead coverage of the beacon antenna. Hence, a beacon station at an airport can assist an aircraft in making its approach.

*Accuracy.*—Range accuracy depends upon radar-sweep speed and the skill of the operator. A reasonable probable of error observation is 0.05 in. on the 5-in. PPI, or 2 per cent of the full-scale sweep. Range error is then $\pm 0.08$ mile on the 4-mile sweep and $\pm 1.8$ miles on the 90-mile and 70- to 160-mile sweeps. Variation of $\pm 0.1$ $\mu$sec in beacon delay increases these possible range errors by $\pm 0.01$ mile.

Azimuth accuracy depends upon the width of the beacon signal and the skill of the operator. It is generally possible to reduce the width of the beacon signals to the half-power beamwidth of 5°. With the aid of the 10° markers provided on the PPI, an average operator can then estimate the center of this 5° signal to $\pm 2°$ or better. On the outer half of the indicator, the azimuth error is likely to be less than $\pm 1°$.

*Traffic Capacity.*—Traffic capacity—the number of radars that can interrogate the beacon simultaneously—can be estimated by the methods discussed in Chap. 6. Assuming that 25 per cent response is adequate for ordinary navigation, we get an estimate of 200 interrogating radars (scanning 360°) when a six-pulse reply code is used. However, interrogation over angles larger than the 5° beamwidth by radars near the beacon station reduces this estimate by an indefinite factor. A maximum traffic capacity of 50 to 100 interrogating radars appears very reasonable under most circumstances.

*Coding.*—Interrogation coding by pulse-width discrimination allows the beacon channel to be independent of the radar channel. For response coding, two two-pulse codes, four three-pulse codes, eight four-pulse codes, 16 five-pulse codes, and 32 six-pulse codes (a total of 62 response codes) are available. Two-pulse and three-pulse codes are used for beacons likely to be subjected to greatest interrogation and hence to highest duty ratio.

*Operational Reliability.*—The system is very reliable. Both radar and beacon are excellently designed and are built of high-quality components. In addition, AN/CPN-6 beacons are usually installed in pairs with automatic alarm and switch-over in case of failure. The simplicity of operation of the AN/APS-10 and its high standard of components and ease of maintenance are fully discussed in *Radar Aids to Navigation*, Vol. 2, Chap. 6, Radiation Laboratory Series.

*Comments.*—Although airborne AN/APS-10's operating with AN/-CPN-6 ground beacons constitute a very satisfactory navigational aid for transport aircraft, a number of major improvements in the system have been suggested. These are briefly discussed below.

*Wavelength*

The optimum wavelength for airborne radar for maximum range and azimuth resolution appears to be about 2 cm; the system described operates at 3 cm. Below 2 cm, absorption by water vapor in the atmosphere begins to be serious for horizon beacon ranges (Sec. 3·6).

*Scatter band*

Precision pretuning of fixed-tuned magnetrons in the manufacturing process has made the scatter band of ± 55 Mc/sec unnecessarily large. Furthermore, tunable magnetrons are now available. A scatter band of ±30 Mc/sec would simplify the beacon-receiver design because it would allow the square-wave-modulated local oscillator and 35 Mc/sec i-f bandpass receiver to be replaced by an unswitched local oscillator and a 60-Mc/sec i-f bandpass receiver (as in the AN/CPN-17 beacon, Sec. 8·12). Tunable magnetrons minimize any possibility of radar interference by allowing a uniform distribution of radar frequencies over the scatter band.

*Range balance*

The existing system is unbalanced, the response link being much stronger (18 db) than the interrogation link. A low-voltage magnetron with 5- to 10-kw pulse power output, similar to that in the air-borne transmitter, would reduce the size and weight of the beacon considerably without any sacrifice in range performance. The resulting loss of 3 to 6 db in the response link is readily allowable. A low-voltage magnetron can easily be designed for a $\frac{1}{2}$ to 1 per cent duty ratio, thus allowing a significant increase in the traffic capacity of the ground beacon.

*Overhead coverage*

Performance of the system when the aircraft is near the beacon (e.g., airport approach) can be improved by providing improved overhead coverage (see Secs. 7·3 and 7·4).

*Simultaneous beacon-radar presentation*

Incorporation of a separate beacon receiver in the airborne radar appears to be worth its cost in weight and complexity. It would allow beacon and radar signals to be observed simultaneously with beacon signals stronger and clearly distinguishable from the radar signals (Sec. 17·3).

**19·4. Lightweight Beacons.**—Three types of lightweight beacon which operate satisfactorily with AN/APS-10 are shown in Fig. 19·3 (AN/UPN-3), Fig. 19·4 (AN/UPN-4), and Fig. 19·5 (an airborne beacon).

AN/UPN-3.—The AN/UPN-3 set (Fig. 19·3) is a lightweight navigational beacon. It has the following characteristics:

Weight (installed): 120 lb, two 60-lb packages
Power supply: 250 watts, 115 volts, 50 to 4000 cps
Antenna: double linear-array; azimuth coverage 360°, elevation half-power beamwidth 10°, gain 10
Transmitter: 2J41 magnetron, r-f pulse power 300 watts, duty ratio 0.3 per cent, 3D21 hard-tube pulser
Receiver: superheterodyne with frequency-modulated local oscillator to cover 110 Mc/sec, sensitivity 90 dbw
Coding: no interrogation coding, 12 response codes
Range: 100 to 150 nautical miles with AN/APS-10

AN/UPN-4.—AN/UPN-4 is a lightweight navigational beacon which can operate from self-contained batteries. It differs from AN/UPN-3 in the following characteristics:

Weight (installed): 55 lb plus 35-lb batteries (for 5-hour operation)
Power supply: 50 watts either from batteries or from 115-volt, 50- to 400-cps line supply

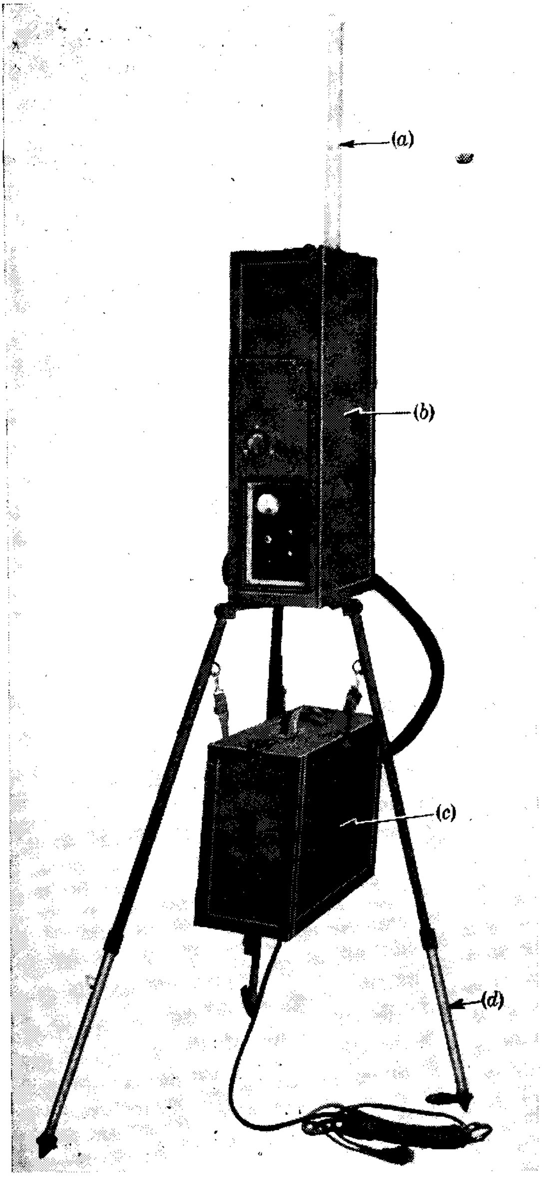


FIG. 19·3.—A lightweight 3-cm ground beacon, AN/UPN-3, set up and ready for operation. (*a*) Antenna. (*b*) Transmitter-receiver. (*c*) Power supply. (*d*) Tripod.

Receiver: crystal-video with sensitivity 77 dbw
Coding: no interrogation coding, 5 response codes
Range: 30 to 50 nautical miles with AN/APS-10

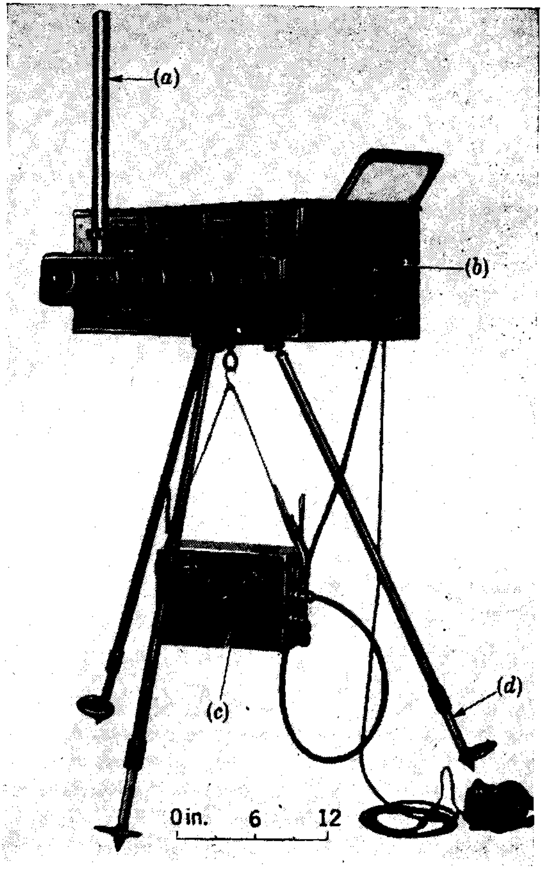


FIG. 19·4.—A lightweight battery-operated 3-cm ground beacon, AN/UPN-4, set up and ready for operation. (*a*) Antenna. (*b*) Transmitter-receiver. (*c*) Battery box. (*d*) Tripod.

An airborne beacon for air rendezvous which operates from an aircraft power supply has been designed. It has the following characteristics:

Weight: 50 lb (single package)
Power supply: 130 watts 115 volts, 400 to 2400 cps; 50 watts, 27 volt d-c

**Antenna**: double linear array with coaxial feed in streamlined housing projecting 1 ft from aircraft
**Coding**: no interrogation coding, 11 response codes

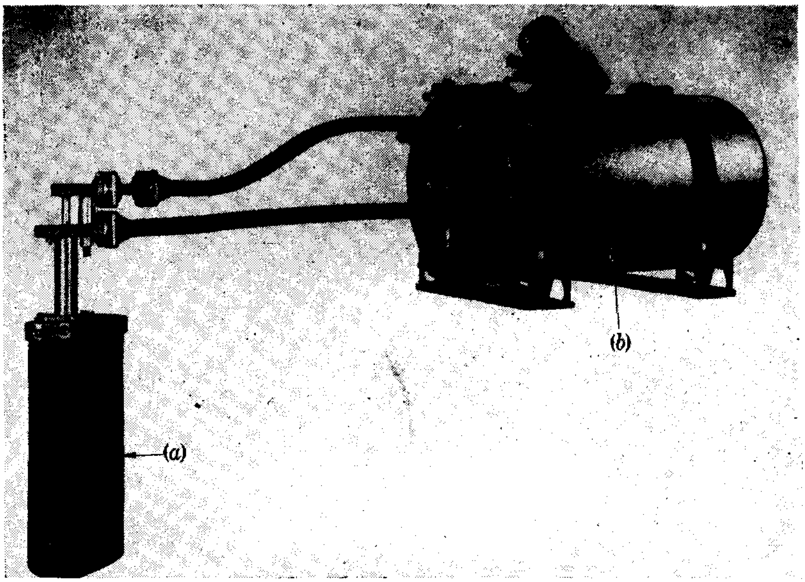


FIG. 19·5.—An airborne 3-cm beacon. The antenna, in a streamlined radome, is connected to the beacon by means of flexible waveguide. A small control box is not shown. (a) Antenna. (b) Transmitter-receiver.

## GROUND-RADAR-AIRBORNE-BEACON SYSTEM

BY G. C. DANIELSON

The description of a ground system operating with the AN/APN-19[1] beacon designed for ground control of all types of aircraft follows. It provides range and azimuth information to the controller on the ground and, by means of an auxiliary radio communication link, to the pilot in the aircraft. Under most circumstances ground control can be accomplished with radar echoes alone; beacons in the aircraft, however, afford at least three important improvements in performance:

1. Range on all aircraft (including small single-engine aircraft) is extended to the horizon for any altitude up to 40,000 ft.
2. Positive simple identification is available for control of heavy air traffic such as that encountered in the vicinity of an airport.
3. Ground and cloud "clutter" are eliminated by the off-frequency beacon response.

---

[1] The AN/APN-19 beacon was affectionately known as "Rosebud" during the war.

**19·5. Ground Radar.**—The ground radar described is designed for long range, accurate azimuth determination, and height finding. A simplified block diagram is shown in Fig. 19·6 and the complete layout on the ground is shown in Fig. 19·7. The important characteristics of the radar as a component of the complete radar-beacon system are listed below.

*General*

Wavelength: 10 cm
Weight (installed): 20 tons, not including power units or shelters
Number of operators: 25 (including communication operators)
Power supply: 40 kva, 120 volt, 60 cps, three-phase
Packaging: six shelters and antenna with tower

*Antenna assembly*

Type: two fan beams, one vertical and one at 45° to vertical; in scanning, the delay between the incidence of the two beams on a target depends upon its height and slant range, and hence provides height information
Vertical beam: 10 by 25 ft paraboloid, three feeds
Slant beam: 10 by 32 ft paraboloid, two feeds
Azimuth coverage: circular scan 360°, 6 rpm, half-power beamwidth 1°
Elevation coverage: 0 to 30° for radar, 0 to 2.2° half-power beamwidth for beacon
Polarization: vertical for vertical beam, 45° for slant beam
Gain (vertical beam): 10,000 for 0 to 2.2° (radar and beacon) 4800 for 2.2 to 8.2° (radar only); 2200 for 8.2 to 30° (radar only)

*Transmitter*

R-f source: 5 HK7 magnetrons
R-f scatter band: 3019 to 2700 Mc/sec for radar; 2965 to 2992 Mc/sec for beacon interrogation
R-f pulse power: 1000 kw (per tube)
Pulser type: rotary spark gap
Pulse-repetition frequency: 350 cps
Pulse duration: 1.0 $\mu$sec

*Beacon receiver*

Type: superheterodyne
Stages: six i-f: two video
I-f pass band: 8 Mc/sec

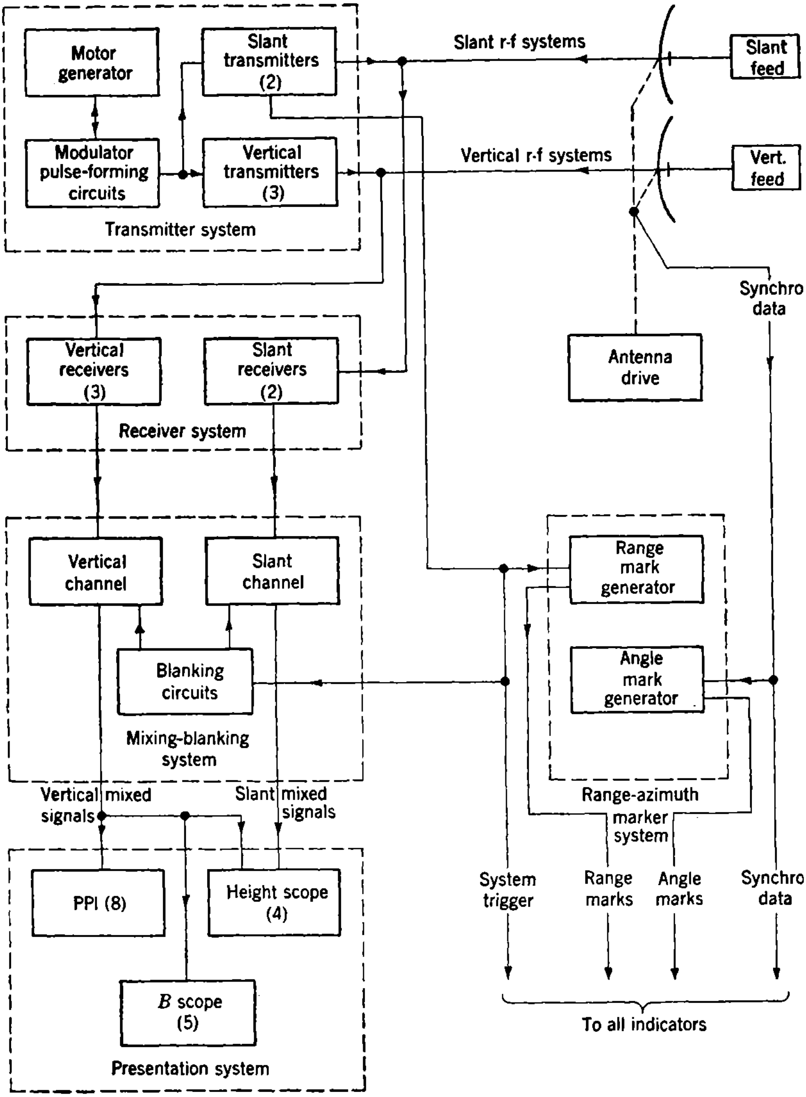


FIG. 19·6.—Simplified block diagram of a 10-cm ground radar set.

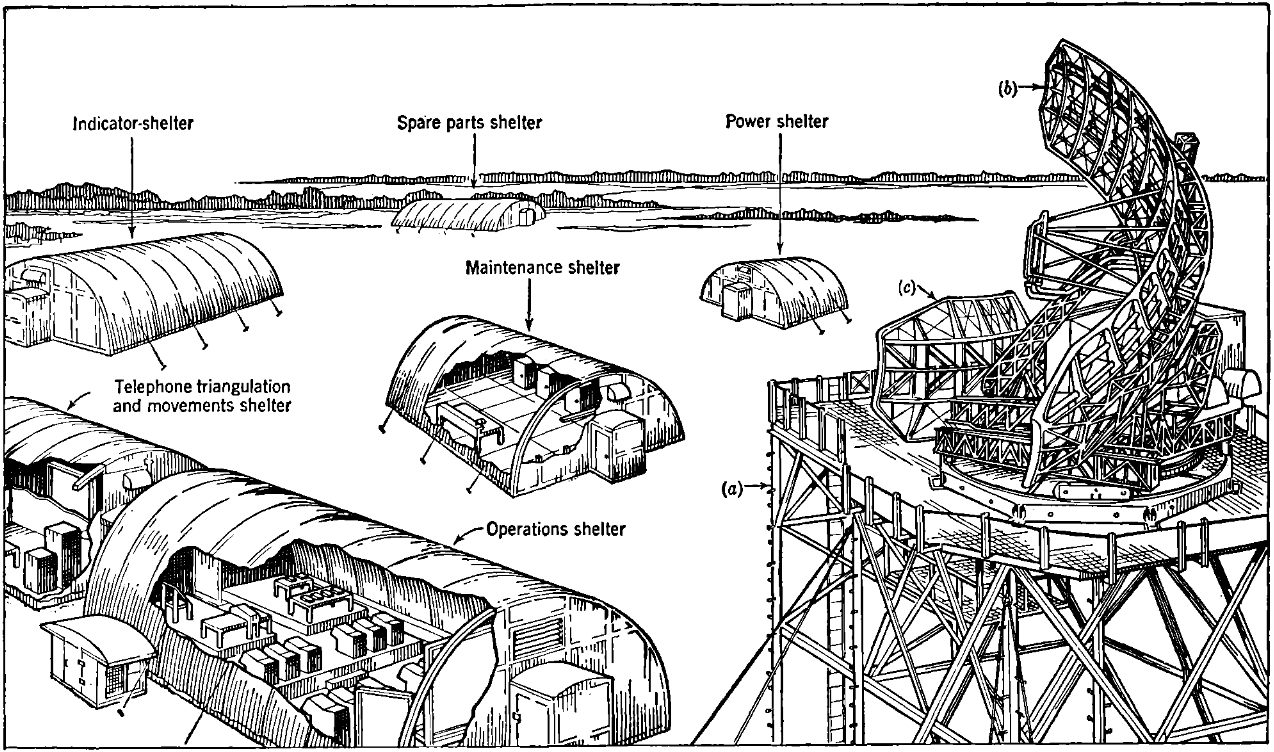


FIG. 19·7.—Ground layout of a 10-cm ground radar military installation. Six shelters are used. (*a*) Antenna mount. (*b*) Slant beam reflector. (*c*) Vertical beam reflector.

Sensitivity (minimum discernible signal): 130 dbw
Special beacon features: beacon local oscillator AFC, separate TR cavity, time-varied gain

*Indication*

Type: eight 12-in. PPI's; five 7-in. B-scopes; four 12-in. height tubes
Sweeps: 10 miles per in., displaying 45 miles; 20 miles per in., displaying 90 miles; 5.5-35 mile per in., continuously variable
Sweep delay: sweep may start anywhere from 0 to 240 miles; off-center operation allows expansion of any sector desired
Range marks: 10-mile range marks
Azimuth marks: 10° angle marks
Selector switch (three-position): allows presentation of radar only, beacon only, or radar and beacon simultaneously

**19·6. Airborne Beacon (AN/APN-19).**—AN/APN-19 is a lightweight airborne beacon which accepts 10-cm interrogation pulses and transmits a 10-cm range-coded reply. It is designed to operate primarily with ground radars. A block diagram of the AN/APN-19 beacon is shown in Fig. 19·8; a photograph of the components is shown in Chap. 16, Fig. 16·3. The characteristics of this beacon are listed below.

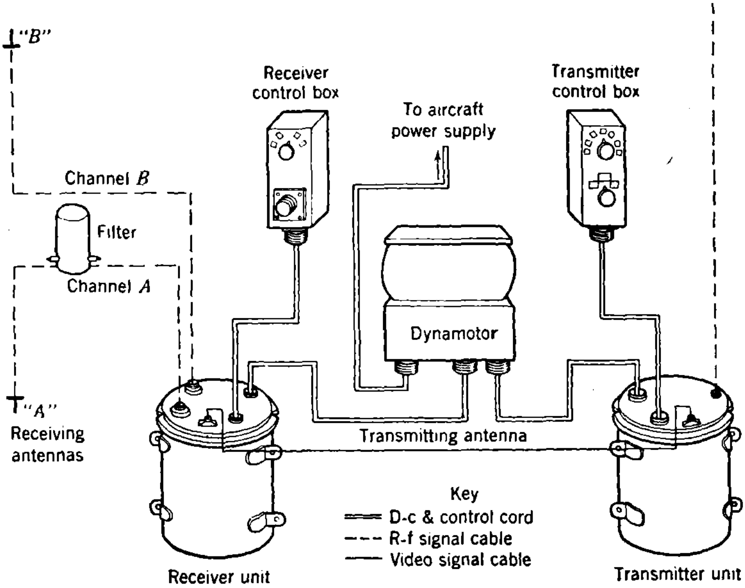


FIG. 19·8.—Block diagram of AN/APN-19. Note two interrogation channels, *A*, *B*, with separate antennas, and the r-f filter used in one channel.

*General*

Wavelength: 10 cm
Weight (installed): 35 lb including cables
Number of operators: one (any crew member)
Power supply: 150 watts; 26.5 ± 2.5 volts d-c
Packaging: receiver (pressurized), transmitter (pressurized), power supply, two control boxes, three antennas (two receiving, one transmitting)

*Antennas*

Type: weatherized vertical quarter-wave dipoles, $\frac{3}{8}$ in. diameter and projecting $1\frac{1}{2}$ in. from aircraft fuselage
Azimuth coverage: 360° unless shielded by aircraft
Vertical coverage: depends upon shape of aircraft surface: typical installation has half-power beamwidth of 50° (+10° to −40° from horizontal)
Polarization: vertical
Gain: 1 to 3 depending upon installation

*Transmitter*

R-f source: triode—2C49 or 2C40 selected for plate pulsing
Response frequency: 2907 Mc/sec
R-f stability: ±3 Mc/sec
R-f pulse power: 100 watts
Maximum duty ratio: 0.4 per cent
Pulser: line-type modulator using three 2D21 gas-filled tubes
Pulse duration: 0.7 $\mu$sec
Blanking-gate duration: 450 $\mu$sec
Overinterrogation protection: blanking gate only

*Receiver*

Type: crystal-video with dual input for allowing two independent receiver channels
Stages: four 7F8 twin triodes
R-f pass band: 300 Mc/sec
Sensitivity (minimum interrogation signal): 77 dbw

*Coding*

Interrogation: r-f filter in one receiver input permits frequency coding in that channel. Four frequencies are used
Response: seven codes (three pulses)

*Other Features*

Power supply: dual output dynamotor giving +360 volts d-c plate supply and −100 volts d-c bias supply; carbon-pile regulator
Controls: off-standby-on switch; response code switch (seven positions); receiver channel switch (2 positions)
Beacon delay: 2.25 ± 0.10 μsec
Monitoring features: test jacks on receiver control box; audio output of interrogation rate available from receiver
Test equipment: AN/APM-53 (includes oscilloscope and signal generator)

**19·7. System Performance.**—The maximum reliable radar search range is 100 miles on small aircraft and 170 miles on large aircraft. For height-finding, the maximum range is 70 miles on small aircraft and 120 miles on large aircraft. Minimum radar range is 3 miles. The accuracy of radar location is ±0.5 miles in range, ±0.5° in azimuth, and ±500 ft in relative height. The performance of the radar with AN/APN-19 beacons, illustrated in Fig. 19·9, is discussed below.

*Coverage.*—Range coverage is limited only by the horizon or by interference nulls. Azimuth coverage is uniform (within 3 db) in all directions from the beacon antenna, unless the antenna is shielded by a portion of the aircraft (for example, by the wings when banking, propellers, or auxiliary fuel tanks). Elevation coverage is limited to 2.2° (half-power beamwidth) from the radar antenna, because only the lower antenna feed of the vertical beam is connected to the receiver. Adequate coverage at short ranges is obtained from side lobes of the lower vertical beam of the radar; elevation coverage appears inadequate, however, for high-flying aircraft at intermediate ranges. Aircraft at 30,000 ft altitude are not within the 2.2° half-power beamwidth when the distance from aircraft to radar is less than 150 miles.

*Beacon Antenna.*—Linear-array feeds of some antennas possess the property of "squint" discussed in Sec. 17·6. Accordingly, if the ground radar has such an antenna, beacon operation is made possible by the addition of an auxiliary beacon-receiving antenna which is so oriented that its beam points in the same direction as the main beam. This makes possible the reception of beacon replies on a frequency different from the interrogating frequency. The gain of the auxiliary antenna is lower than that of the main beam, but there is sufficient power margin available so that beacon operation to horizon range is assured.

*Range.*—Maximum reliable beacon range is at least 270 statute miles, if not limited by the horizon. Theoretical maximum range calculations, computed as in the example in Sec. 3·11, indicate (1) that the interrogation link is appreciably stronger (6 db) than the response link, and (2) that the reliable response range is about 800 statute miles. At an actual maximum range of 270 miles, approximately 9 db are thus available for

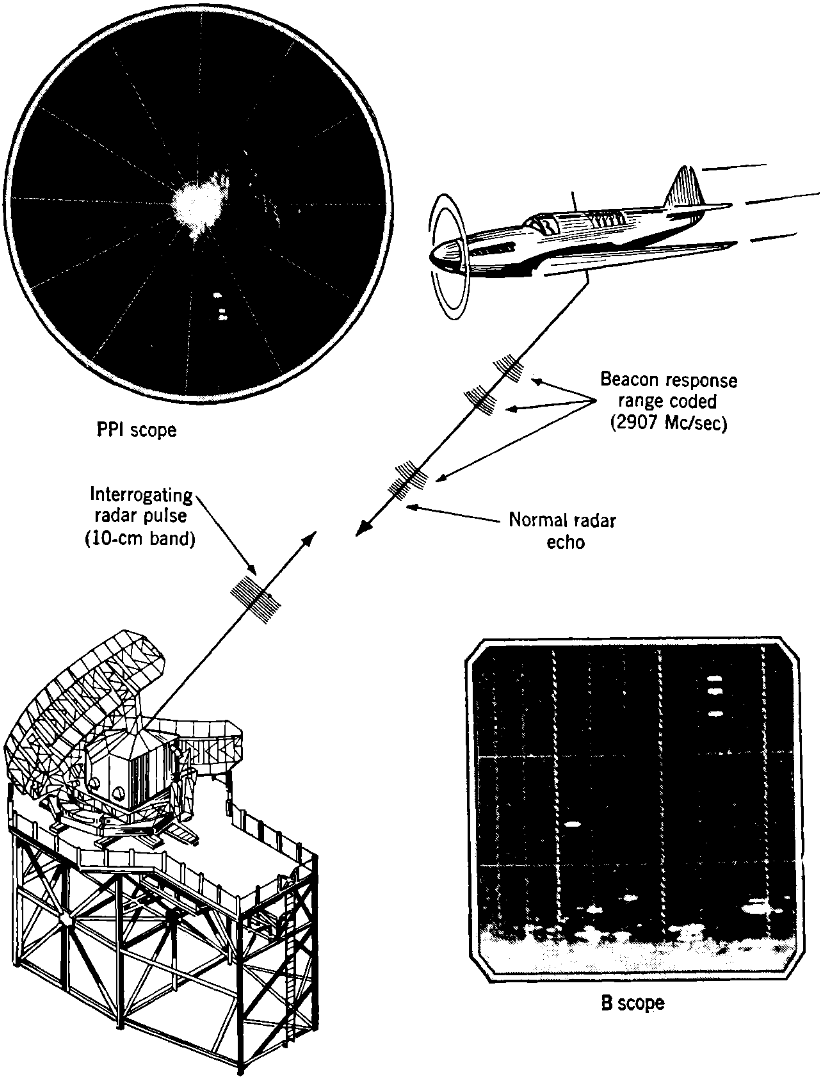


FIG. 19·9.—Beacon operation and target appearance on indicators of the large 10-cm ground radar.

overcoming possible interference nulls. The radar described is generally free from interference effects, partly because of its antenna design and partly because of the way it is usually sited.

Minimum range for azimuth discrimination, even with the time-varied gain feature, is 5 to 10 miles. Hence, for airport control, it is advantageous to locate the ground radar at least 10 miles from the airport.

*Accuracy.*—Assuming a probable observation error of 0.05 in., the range error is ±1.0 mile on the 20-mile/in. sweep, ±0.5 mile on the 10-mile/in. sweep, and ±0.3 mile on the 6-mile/in. variable sweep speed. Variation in beacon delay is negligible.

The azimuth error for 1° signals (the half-power beamwidth) is approximately $\frac{1}{20}$ the distance between 10° markers over the central and outer portion of the PPI. This represents a probable azimuth error of about ±0.5°.

Height information is obtained only by radar echoes. The probable error in relative height is about ±500 ft.

The maximum accuracy in range, azimuth, and height is inherently limited by the slow rotation speed (6 rpm) of the large antenna. Special computing devices designed to predict the exact location of the aircraft between scans can, however, reduce this difficulty.

*Traffic Capacity.*—Since it is practically impossible to overload the airborne beacon with scanning ground radars, the traffic capacity of the system is limited only by the operational capacity of the ground equipment. Allowing two observations per PPI, the normal complement of eight tubes permits plotting and following 16 aircraft (or 16 groups of aircraft) simultaneously. An alternative to providing more indicator tubes is to photograph each complete scan and project it within 10 sec on a vertical plotting board. Such additional equipment, which allows 50 or more aircraft to be numbered, identified, and followed on one large, illuminated, semitransparent screen, is particularly valuable for control of heavy airport traffic.

*Coding.*—Since pulse-width discrimination is omitted from the beacon in order to lessen weight, size, and power, all 10-cm ground radars can interrogate the beacon on every scan. The absence of interrogation coding causes no difficulty at the beacon because of the very low duty ratio resulting from a scanning ground radar, and no difficulty at the radar because the three-position selector switch on each PPI can separate beacon and radar presentations.

A total of seven response codes is available—one one-pulse code, two two-pulse codes, and four three-pulse codes.

*Operational Reliability.*—The system has a high degree of reliability. The ground radar is a high-performance device maintained at peak performance by trained crews. The airborne beacon is built to withstand large variations in ambient temperature and pressure. Special transmitter cavities give good frequency stability over these wide variations in ambient conditions. A serious problem with some aircraft, however, is the difficulty of determining a location for the beacon antennas which will provide adequate coverage in all aircraft aspects during level flight and during normal climbs, dives, and banks.

**19·8. Comments.**—The radar-beacon system described is one of the best currently available systems of air traffic surveillance and control for most types of aircraft. It provides particularly valuable navigational assistance to single-place aircraft, which are too small to carry complete radar equipment but are large enough to carry a 35-lb beacon. It does not include airport approach and landing facilities. Some suggestions that have been made to increase the usefulness of the system are discussed briefly below.

*Antenna Coverage.*—The elevation coverage of the ground radar for beacon is nominally 0 to 2.2°; this hardly seems adequate. By using the middle antenna feed (double 60° horn, 2.2° to 8.2°, gain 4800) as well as the lower antenna feed, the coverage at intermediate ranges and high altitudes becomes excellent. The improved performance seems well worth the cost of an additional beacon receiver in the radar.

The quarter-wave dipole beacon antennas, projecting $1\frac{1}{2}$ in. from the fuselage, provide good coverage for most aircraft but are entirely unsuitable for some aircraft (for example, the P-38). For these difficult cases, a streamlined two-element vertically polarized linear-array antenna has been developed. This antenna is built to project about 8 in. from the fuselage, instead of only $1\frac{1}{2}$ in., and is thus more suitable for aircraft with flat bottoms and low wings.

*Range Balance.*—The system is slightly unbalanced, the interrogating link being about 6 db stronger than the response link. In general, it is better to have the response link somewhat stronger so that the beacon signals will always appear strong. In this system, the strength of the interrogation link would be reduced, and at the same time protection against beacon-receiver crystal damage would be increased, if slightly less sensitive high-burnout crystals were used in the beacon receiver. The usefulness of such a change is, however, questionable.

*Beacon Delay.*—The constant beacon delay of 2.25 μsec (0.2 mile), when significant, is subtracted from a beacon-range reading. Compensation for this constant error can be accomplished more easily by delaying the output of the radar receivers by 2.25 μsec. If desired, the magnitude of the beacon delay can be reduced considerably.

*Coding.*—For ground control of heavy air traffic, several methods of increasing the amount of information conveyed by response coding have been suggested. One of the more elaborate methods suggests six-pulse codes. The first pulse gives range, the second gives altitude if the spacing between first and second pulses is controlled by the aircraft altimeter, and the remaining four pulses give 10,000 identification codes by allowing each space ten values. Special decoding equipment would be needed at the ground radar (see Sec. 5·15).

**19·9. Interrogation Channels.**—AN/APN-19 can also be used with automatic-tracking ground radars, such as the SCR-584 (see Fig. 19·10), if interrogation channels are provided to prevent beacon "stealing."[1] The simplest type of interrogation channeling is by frequency. AN/APN-19 is already provided with a separate receiving antenna, an adjustable r-f filter cavity, and a separate detector and first stage in the receiver. By switching from one receiver input to the other it is possible to change from wideband reception (for search) to narrowband reception (for tracking). For aircraft installations requiring the larger linear-array beacon antenna, it is preferable to use an r-f antenna switch instead of the additional receiving antenna in order to minimize drag.

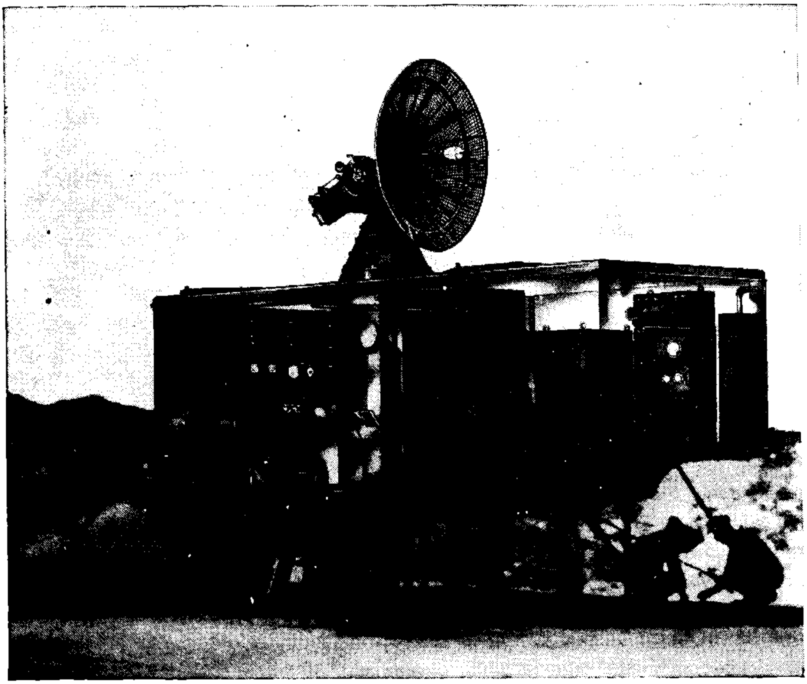


FIG. 19·10.—The SCR-584 automatic-tracking radar. Used with the AN/APN-19, it provides very accurate control of aircraft at distances up to 100 miles.

A more effective and flexible solution to the problem of getting more channels is offered by double-pulse interrogation coding (Sec. 5·5). The

---

[1] See Sec. 6·1.

ground-tracking radar interrogates the beacon with a double pulse of 3-, 6-, 9-, or 12-μsec separation. The beacon receiver is followed by a decoder which allows only the appropriate double pulse to interrogate the beacon. Selection of any channel is accomplished in the air by a simple electrical four-position switch. This improvement is incorporated in the AN/APN-19A.

## SHIP INTERROGATOR-RESPONSOR–SHORE-BEACON SYSTEM

BY P. A. DE PAOLO

**19·10. Introduction.**—In the following pages, we describe a hypothetical ship interrogator-responsor–shore-beacon system suitable for off-shore navigation of surface vessels. By including this description, we do not mean to imply that the system described is necessarily the best that can now be designed for the purpose. In fact, a microwave system offers a number of advantages over the system to be described.[1] However, we wish to include a description of a long-wave navigational system using lobe switching for azimuth determination. The system described purports to be an improvement over many such systems that have been used.

The system described below gives line-of-sight range (or a maximum reliable overwater range) of 75 miles. Azimuth data are obtained by using a lobe switched antenna system on the interrogator-responsor. The system uses off-frequency response in order to eliminate ground clutter when ships are navigating close to land masses. The maximum range that can be obtained with this system depends on the heights above sea level of the beacon antenna and the interrogator antenna.

Navigation is accomplished either by homing on a particular beacon, by taking fixes on one or more beacons, or by triangulation with respect to several beacons. Any of these methods requires that the navigator have accurate maps or charts of the area to be traversed and that the location of the beacons in this area be known accurately.

**19·11. Interrogator-responsor.**—Figure 19·11 is a block diagram of the interrogator-responsor and its display and control circuits. The system is divided into three packages: the antenna and lobe switch; the transmitter and receiver; and the indicator central and ranging unit.

Timing and synchronizing are accomplished in this system by means of a crystal-controlled oscillator operating at a frequency of 81,882 cps. The period of this oscillator is equal to 12.3639 μsec, and corresponds to the time required for an r-f pulse to travel a round-trip nautical mile. The sine-wave output of the oscillator is squared, formed into pips, and

---

[1] For a description of a proposed microwave ship radar with beacon navigation facilities, see *Radar Aids to Navigation*, Vol. 2, Secs. 10·2 to 10·5, Radiation Laboratory Series.

amplified. The repetition rate of the amplified output pips is first divided by 2, then by 5, and finally by 30. The original pips give accurate range marks for 1-mile intervals; the first and second divisions give range marks for 2- and 10-mile intervals; the output pulses of the last divider have a PRF of 81,882/300, or 269.6 cps; after suitable amplification they are used to trigger the modulator and the sweep circuits on the indicators.

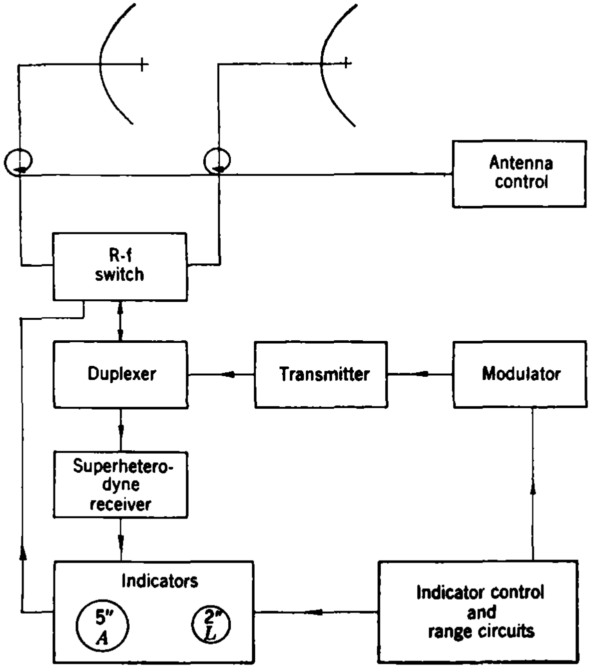


FIG. 9·11.—Block diagram for a shipborne interrogator-responsor. The antennas are Yagi arrays with a gain of 6, switched 60 times per sec. The receiver has a sensitivity of 123 dbw. The transmitter has a pulse power output of 750 watts. The display uses a 5-in. A-scope for range data and a 2-in. L-scope for pip-matching. The range scope has sweeps of 1, 2, 5, 10, 20, and 40 nautical miles with range marks of 1, 5, and 10 nautical miles. The transmitter and receiver tune from 480 to 520 Mc/sec.

A random variation of the division ratio of the last frequency divider is intentionally introduced. This varies the pulse-repetition frequency somewhat, without introducing any error in range, and prevents beacon stealing (Sec. 6·1).

The indicators are of conventional design and include a 5-in. A-scope for ranging and a 2-in. L-scope for pip matching. The 5-in. range scope has a total of five sweep speeds, corresponding to total sweep lengths of 1, 5, 10, 20, and 40 nautical miles. The range marks are mixed with the output of the responsor and displayed simultaneously on the range scope.

The L-scope sweep is initiated by the trailing edge of the pulse from a continuously variable delay multivibrator. The sweep length is equal to 1 nautical mile. The delay multivibrator is adjusted so that the desired beacon signal appears on the L-scope. The input signals to the L-scope are switched at 60 cps in synchronism with the antenna switch so that the beacon signal appears on the screen alternately as a deflection to the left and then as a deflection to the right. Azimuth data are obtained by rotating the antenna until the left and right signals have equal amplitudes, indicating that the beacon is being received with equal signal strength on both antennas, as shown in Sec. 1·1, Fig. 1·6. It is important that the receiver gain be so adjusted that the receiver does not saturate. Unless this is done, both signals appearing on the L-scope will have equal amplitudes over a wide arc, and incorrect azimuth data will be obtained.

The rest of the interrogator-responsor is of simple but reliable design. The modulator uses a hydrogen thyratron with resonance charging and a holding diode. This will allow the use of a low-voltage power supply for the modulator.

The transmitter uses a 15E tube in a half-wave coaxial-line oscillator. The main requirements for the transmitter are that it deliver adequate power over the operating range of frequencies and that it have good frequency stability.

The receiver is of the superheterodyne type. The first r-f stage is a grounded-grid amplifier which is used to improve the noise figure of the receiver. The i-f and video systems are of conventional design. The receiver is equipped with time-varied gain in order to prevent blocking when navigating close to a beacon. In addition, the receiver is provided with a manual gain control for use during pip matching.

The antenna system consists of two Yagi arrays mounted on a cross-arm. The antenna system is so arranged that it may be rotated 360° in azimuth. The position of the antenna may be controlled by means of a simple motor drive or by a manual control. The direction in which the antenna is pointing is indicated by means of a selsyn motor at the operating position.

*Summary.*—The characteristics of the interrogator-responsor are listed below.

*General*

Frequency: 480 to 520 Mc/sec
Weight (installed): about 150 lb
Number of operators: one

Power supply: 400 watts, 115 five, 60 cps
Packaging: 3 units—transmitter-receiver; indicator, indicator control, and antenna control panel; antenna assembly

*Antenna Assembly*

Type: two separate Yagi arrays with r-f switch
Azimuth coverage: 360°; half-power beamwidth for each antenna 60°
Polarization: vertical
Gain: 6

*Transmitter*

R-f source: one 15E triode
Frequency range: 480 to 520 Mc/sec
R-f pulse power: 750 watts
Pulser type: 3C45 hydrogen thyratron
Pulse-repetition frequency: 270 cps
Pulse duration: 0.75 $\mu$sec

*Receiver*

Type: superheterodyne
Stages: one r-f, six i-f, two video
I-f bandwidth: 2.0 Mc/sec
Sensitivity: 113 dbw
Special features: time-varied gain, remote gain control

*Indication*

Type: 5-in. A-scope; 2-in. L-scope for pip matching
Sweeps: 1-, 5-, 10-, 20-, and 40-nautical-mile
Range marks: 1-, 2-, and 5-mile

**19·12. The Ground Beacon.**—Figure 19·12 is a block diagram of the ground beacon used in this system. The receiver uses a single r-f stage ahead of the mixer. The r-f amplifier is of the grounded-grid type. The receiver should be designed to have good frequency stability and it should not block when receiving interrogations at short ranges.

The output signals of the receiver are fed into the coder. The coder, in combination with the modulator, provides 10 reply codes. These codes are selected by means of the code switch. The coding is a combination of pulse-width and multiple-pulse coding.

The modulator uses thyratrons in order to cut down the stand-by power. The multiple pulses are generated by using two thyratrons and two pulse-forming networks. The pulse-width coding is accomplished by changing the constants of the pulse-forming network.

The coaxial-line transmitter uses a miniature tube (6F4) which is operated well below its maximum ratings to ensure reliability and to minimize frequency drift due to variation of duty ratio. The transmitter must have excellent frequency stability and must deliver adequate power over the frequency band of operation.

The antenna is a half-wave, center-fed dipole, essentially omnidirectional, and matched from 480 to 520 Mc/sec. The transmitter and receiver are duplexed into one antenna.

Power is supplied from batteries that can be charged by energy from a local power line, from an automatic motor-generator set, or from a wind charger. The method used will depend on local conditions.

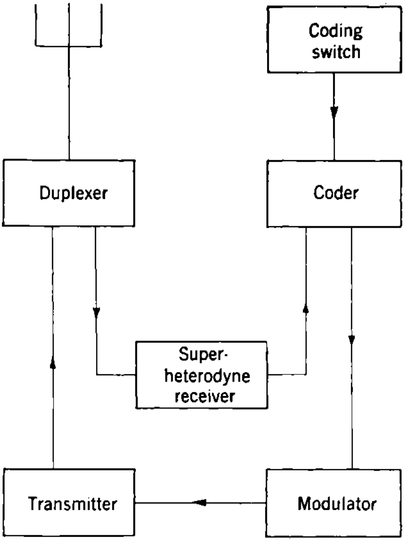


FIG. 19·12.—Block diagram of the shore beacon. This beacon has a receiver sensitivity of 110 dbw and a power output of 100 watts. It provides for 10 reply codes. The receiver and transmitter are duplexed into a common antenna and may be tuned from 480 to 520 Mc/sec.

*Summary.*—The characteristics of the ground beacon are listed below.

*General*

Frequency: 480 to 520 Mc/sec
Weight (installed): about 40 lb (does not include batteries or charger)
Number of operators: none
Power supply: batteries
R-f lines: flexible coaxial RG-8/U
Packaging: three units—antenna assembly; transponder; and power supply (control box is built into the transponder)

*Antenna*

Type: single, center-fed, half-wave dipole
Coverage: 360° in azimuth; can be modified to suit local needs
Polarization: vertical
Gain: 2

*Transmitter*

R-f source: one 6F4 tube
Frequency range: 480 to 520 Mc/sec
R-f pulse power: 100 watts
Pulser type: (2) 2D21 thyratron
Pulse rate: 0-5000 pps
Pulse duration: 0.75 or 1.5 μsec

*Receiver*

Type: superheterodyne
Number of stages: one r-f (grounded-grid), six i-f, two video
I-f bandwidth: 2.0 Mc/sec
Sensitivity: 105 dbw
Special features: provision to prevent c-w blocking

*Coding*

Interrogation: none
Reply: range- and pulse-width coding with two pulse widths (0.75 and 1.5 μsec) and two pulse spacings (3 and 5 μsec)
Number of possible codes: 10

**19·13. System Performance.**—Because the system described above has never been built or used, no actual data on its performance can be given. From experience with similar systems, however, predictions as to the probable performance of this system can be made.

*Coverage.*—The range coverage of this system will be limited by the horizon or, rarely, by interference nulls. The azimuth coverage is 360°. However, for any given antenna setting, beacons located over at least a 90° arc will be interrogated. This is undesirable but very difficult to overcome unless antennas of higher resolution are used; it is one of the arguments mentioned above for using microwaves.

*Range.*—The computed reliable free-space range for this system is about 75 nautical miles. This will not be the usable range because of the horizon limitation and because the system works over water. If the beacon is located on high terrain (about 200 ft above sea level) and the interrogator antenna is mounted at about 25 ft above sea level, the computed overwater range (from Eq. (4) Sec. 4·1) is about 60 miles. Since the horizon range is only 22 miles for antennas at these heights a reliable horizon range of about 25 nautical miles can reasonably be expected.

The system uses off-frequency response; hence the minimum range is limited by the pulse width of the interrogator, and will be approximately $\frac{1}{10}$ mile. Responses picked up by the antenna side lobes should not appear if the responsor is operated at low gain at close ranges.

*Accuracy.*—If it is assumed that the total usable sweep length on the 5-in. A-scope is 4 in., and that the probable observation error is 0.05 in., then the range error is ±0.01 mile on the 1-mile sweep, ±0.02 mile on the 2-mile sweep, ±0.06 mile on the 5-mile sweep, ±0.12 mile on the 10-mile sweep, ±0.25 mile on the 20-mile sweep, and ±0.5 mile on the 40-mile sweep. It is assumed that the beacon delay is small and of constant value.

The azimuth accuracy of this system will depend almost entirely on the original alignment of the system and the skill of the operator. Several radar systems with switched antennas at 500 Mc/sec have been used. The azimuth accuracy of these systems was about ±1.0° to ±2.0°. It is reasonable to assume the same degree of accuracy for the system described. It is expected that the azimuth accuracy will decrease when the vessel is sailing in heavy seas, because the beacon signal will vary with the pitch and roll of the vessel and more time will be required to get a fix under these difficult conditions.

*Traffic Capacity.*—The system described has a high enough safety factor in both links so that the beacon signal should be very strong at all times. The operator should be able to get good fixes, therefore, even when the beacon is counting down due to overinterrogation. Assuming that 25 per cent response is adequate, we find that about 100 interrogators can see the beacon at any one time. This figure, of course, includes interrogators which do not use the beacon intentionally but which interrogate it nevertheless because of their wide antenna patterns. Saturation of the response link is likely to limit the usefulness of the system because of "fruit" on the display. This limitation, which occurs at a traffic level that depends on the expansion of sweep used, will probably occur at a density appreciably lower than that given above for the interrogation link.

If the shore beacon system were to be laid out carefully, and several different beacon frequencies used, the traffic capacity of the system would be increased manyfold.

*Comments.*—The ship-interrogator-responsor–shore-beacon system described should give good results as a navigation system. In combination with a small radar for collision warning, it should constitute an excellent navigation aid for coastwise shipping.[1] The receiver and transmitter of the interrogator-responsor may be tuned to the same frequency and used as a radar system, which should give a radar range of 5 miles or better on a large vessel; radar performance will obviously be poor.

---

[1] In contrast, the microwave ship-navigation radar described in *Radar Aids to Navigation*, Vol. 2, Chap. 10, Radiation Laboratory Series, provides both radar and beacon information.

## SHORAN–A PRECISION BEACON-NAVIGATION SYSTEM

By G. C. Danielson

A typical example of a precision beacon system is the Shoran system, produced (except for the computer, which is omitted from this discussion) by RCA Laboratories under the sponsorship of the Wright Field Radio and Radar Laboratory. Shoran is an $H$-navigational system which uses an AN/APN-3 interrogator-responsor in the aircraft, operating with two AN/CPN-2 beacons on the ground. The navigator in the aircraft determines his position accurately by measuring range only to the two fixed beacons at known locations on the ground. The system is designed for any application requiring precision range measurement, such as aerial mapping. A block diagram of the Shoran system is shown in Fig. 19·13.

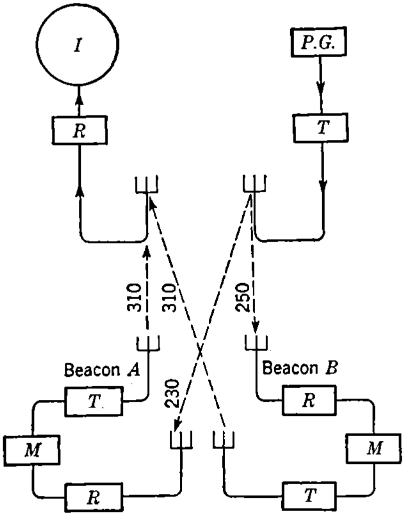


FIG. 19·13.—Block diagram of the Shoran navigational system. Figures refer to frequencies in Mc/sec. $R$—Receiver. $T$—Transmitter. $M$—Modulator. $I$—Indicator. $P.G.$—Pulse generator.

**19·14. Interrogator-responsor (AN/APN-3).**—The AN/APN-3 is designed solely for operation with ground beacons in an H-type system. It indicates accurately ranges to two beacons and is not required to indicate either radar echoes[1] or beacon azimuth. A photograph of the Shoran airborne equipment is shown in Figs. 19·14 and 19·15. Characteristics of the AN/APN-3 interrogator-responsor are listed below.

*General*

Wavelength: 1 m

Weight (installed): 205 lb, including cables but not bombing attachments; 280 lb with bombing attachments and inverter power supply

Number of operators: one

Power supply: 459 watts, 27.5 volts d-c; 665 volt-amp, 115 volts, 400 to 2400 cps

Packaging: four units; transmitter, timer-indicator, comparator, two antennas; bombing attachments include computer, comparator, and two PDI (pilot direction indicators)

---

[1] Ground echoes are sometimes used for altitude measurement.

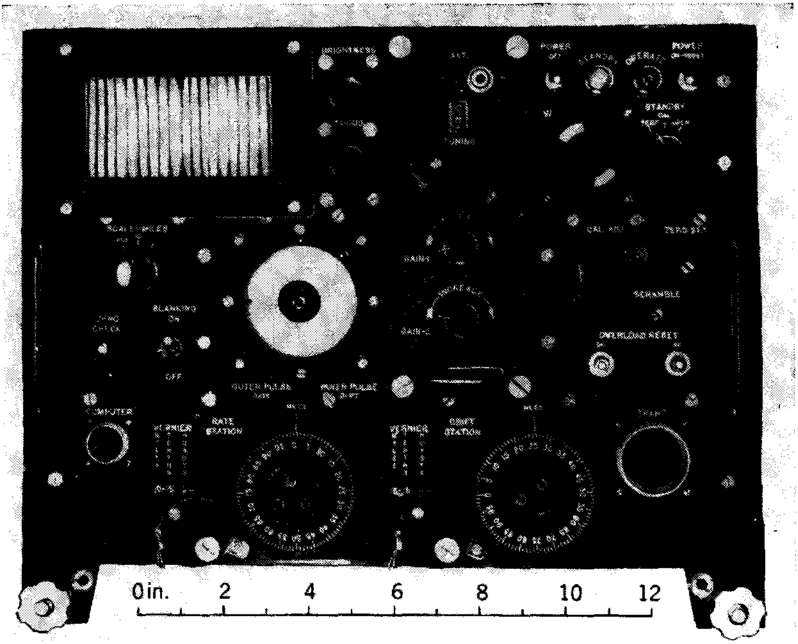


FIG. 19·14.—Timer-indicator unit of the Shoran airborne interrogator-responsor (AN/APN-3).

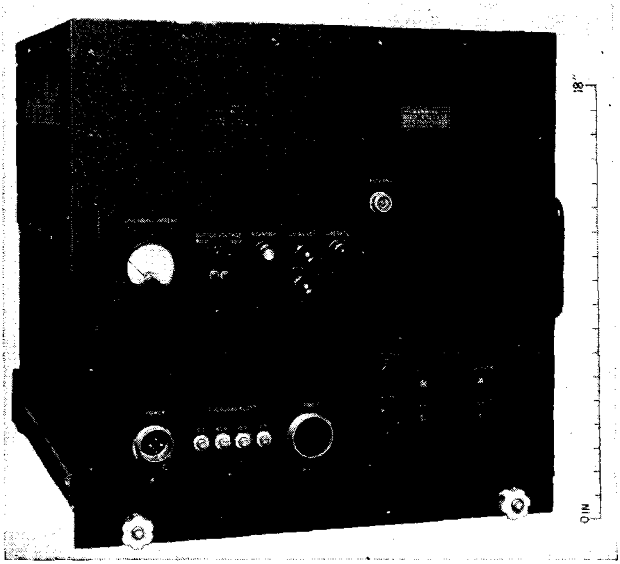


FIG. 19·15.—Transmitter of Shoran airborne interrogator-responsor (AN/APN-3).

*Antenna Assembly*

Type: vertical coaxial dipole; separate receiving and transmitting antennas; one antenna mounted on top, the other on bottom of aircraft
Azimuth coverage: 360°
Elevation coverage: ±45°
Polarization: vertical
Gain: 1·5

*Transmitter*

R-f source: triode, (2) 4C28
Radio frequency: two frequencies (one for each beacon) separated about 20 Mc/sec in the 220- to 260-Mc/sec band.
R-f stability: 0.2 per cent
R-f pulse power: 12 kw
Frequency switching rate: 10 cps ($\frac{1}{30}$ sec of interrogation for each beacon separated by $\frac{1}{60}$-sec intervals of silence)
Pulser type: hard-tube, (2) 3E29
Pulse rate: 931 pps derived from 93-kc master oscillator
Pulse duration: 0.5 μsec

*Receiver*

Type: superheterodyne
Frequency range: 220 to 330 Mc/sec
Stages: six i-f
I-f pass band: 4 Mc/sec
Sensitivity: 128 dbw

*Indication*

Type: (1) distance dials; (2) two PDI (pilot direction indicators); one PDI indicates displacement from track (full-scale deflection either 1000 or 400 ft) and the other PDI indicates rate of approach to track; (3) one 3-in. J-scope
Sweeps (J-scope): 1-mile time base (7 in./mile); 10-mile time base (0.7 in./mile); 100-mile time base (0.07 in./mile).
Range measurements: beacon range corresponds to phase shift in precision goniometers (100 mile, 10 mile, 1 mile) and is indicated on Veeder-root counter; accuracy of 1-mile goniometer is 1.5° or 22 ft; smallest dial on counter indicates 0.01 mile allowing estimates to 0.001 mile

**19·15. Ground Beacons (AN/CPN-2).**—Two AN/CPN-2 beacons are required for operation with the AN/APN-3 interrogator-responsor.

In the usual application, the aircraft flies a circular track around one of the beacons (called the "drift station") and determines its position on this circular track by the other beacon (called the "rate station"). The beacons are interrogated on different frequencies but respond on a common frequency. Photographs of the Shoran ground beacon are shown in Figs. 19·16, 19·17, and 19·18.

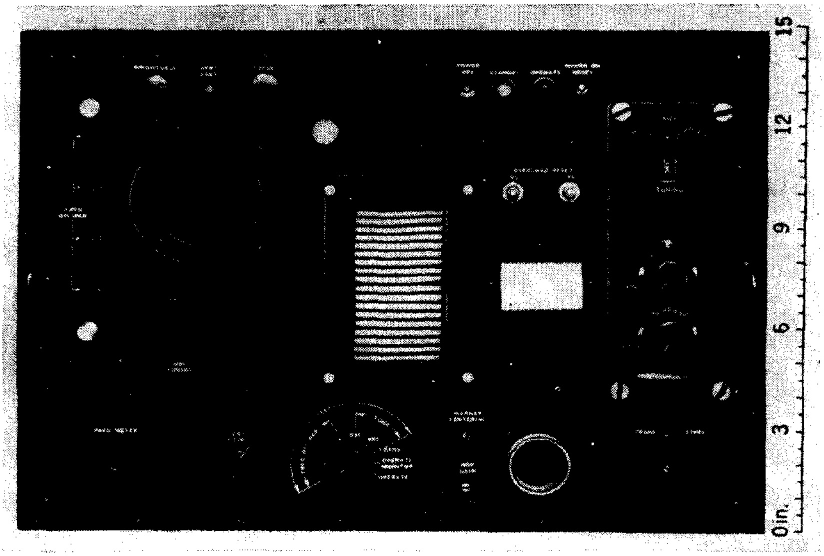


FIG. 19·16.—Shoran beacon (AN/CPN-2) receiver-monitor unit.

Characteristics of AN/CPN-2 are as follows.

*General*

Wavelength: one meter
Weight (installed): 1150 lb, including 310 lb for beacon, 460 lb for antenna and mast, and 380 lb for two power units and fuel cans
Number of operators: one
Power supply: 0.4 kw 24-volts d-c, 1.2 kw, 115 volts, 400 cps

*Antenna assembly*

Type: vertical coaxial dipole set in a corner reflector; separate receiving and transmitting antennas mounted on a 50-ft mast
Azimuth coverage: 90°
Elevation coverage: ±45°

Polarization: vertical
Gain: 6

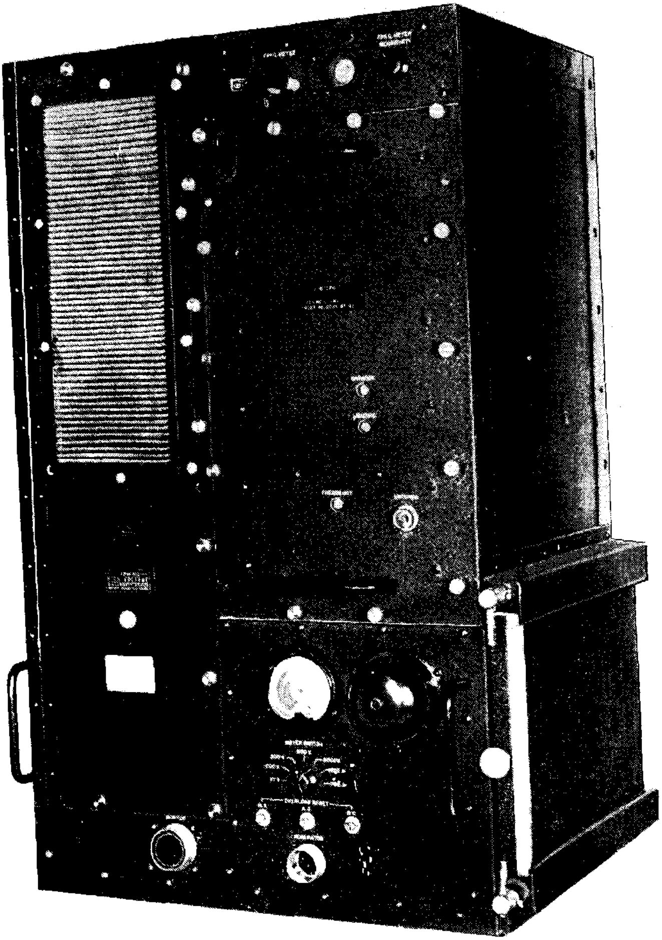


FIG. 19·17. Shoran beacon transmitter unit.

*Transmitter*

R-f source: triodes, (2) 4C28
R-f response frequency: selected in 290 to 320-Mc/sec band b tunable transmitter
R-f frequency stability: 0.2 per cent

R-f pulse power: 30 kw
Maximum duty ratio: 0.4 per cent
Pulser type: hard-tube, (4) 3E29
Pulse duration: 0.55 μsec
Blanking gate duration: 5 μsec
Overinterrogation protection: high-voltage overload relay and warning bell

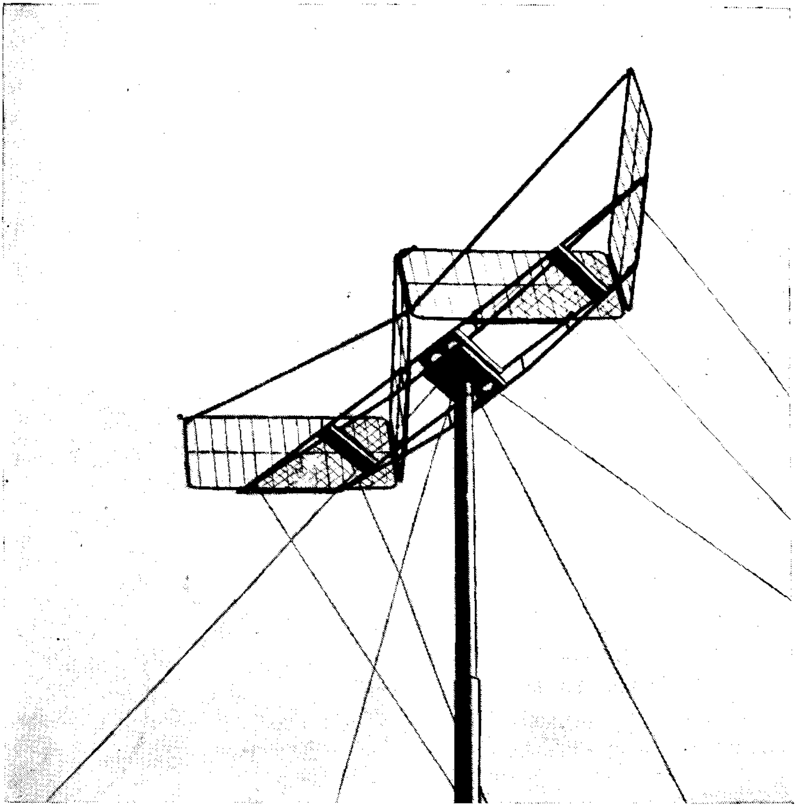


FIG. 19·18.—Shoran ground beacon (AN/CPN-2) antenna installation.

*Receiver*

Type: superheterodyne (identical with AN/APN-3 receiver)
Frequency range: 220 to 330 Mc/sec
Stages: six i-f
I-f passband: 4 Mc/sec
Sensitivity: 128 dbw

*Coding*

Interrogation: frequency discrimination by narrow-band receiver

Response: no range coding; provisions in aircraft to identify responses to its own interrogations and to distinguish drift from rate station (one appears as outward, the other as inward pip on circular trace in J-scope)

*Special features*

Power units: two PU-4/CPN-2 gasoline-power units are supplied with each beacon; each unit is 24-volt d-c, 115-volt 400-cps, 1.7-kw supply.

Frequency monitor: provisions for monitoring both interrogation and response frequencies

Beacon delay monitor: 3-in. A-scope with 1-mile sweep allows monitoring of delay to better than ±0.1 μsec.

Calibration broadcast: pulses from a standard temperature-controlled crystal (93,109 ± 2 cycles) are broadcast by ground station at intervals of exactly 100 statute miles.

These pulses are used to check the time base in the airborne equipment.

**19·16. System Performance.** *Coverage.*—Range coverage is limited only by the horizon. Azimuth coverage is limited to a 90° sector from each beacon station when the corner reflectors on the beacon antenna are used; azimuth coverage is uniform in all directions when the corner reflectors are removed. Elevation coverage is uniform from 0 to 45°. Hence, when airborne and beacon antennas are properly mounted, the system gives essentially complete line-of-sight coverage.

*Range.*—Maximum reliable range is limited only by the horizon (about 280 miles at 40,000 ft). Theoretical maximum range calculations indicate adequate reserve power for strong signals (20 db above minimum discernible) and for interference nulls.

*Accuracy.*—The accuracy of the Shoran system depends as much upon external factors, such as variations in propagation velocity, as upon instrumental errors. By adjustment of the timing crystal with the aid of standard frequency transmissions the crystal error can be made negligible. With careful adjustment the errors in the equipment can be reduced to ±25 ft. at any operational distance. This precision and the speed with which fixes may be taken make the system the best existing one for precise navigation for any purpose.

*Traffic Capacity.*—Twenty interrogator-responsors can operate with one pair of ground beacons simultaneously.

*Coding.*—Interrogation coding is by frequency channeling only and there is no response coding. Provisions are made in the interrogator-responsor to distinguish the constant-range (drift, or "cat") station from the variable-range (rate, or "mouse") station.

*Operational Reliability.*—The equipment is designed for just one purpose: precision navigation as an *H*-system. The operation of the airborne equipment can be made almost entirely automatic; course information can be delivered to the automatic pilot and fixes can be taken automatically. Hence the period of training required to operate the equipment is very short and the possibility of personnel errors is reduced to a minimum.

*Comments.*—Shoran is the most accurate radar system for aircraft navigation to horizon ranges that has been developed to date (1945). It owes its great precision to the inherent advantage of requiring range measurements only (because these can be determined much more precisely than azimuth- or elevation-angle measurements) and to the excellence of its instrumentation, which utilizes this inherent advantage by employing devices such as expanded sweeps, tracking circuits, calibration checks, and automatic operation.

Shoran is not adapted to any applications other than those using *H*-navigation, since the equipment is not designed to indicate either radar echoes or beacon azimuth. Although its dipole airborne antenna has the advantage of low drag on the aircraft, it prevents getting the large traffic capacity associated with scanning radars even when range-coded beacons are used with the latter.

Extensive experimental work has shown that it is possible to incorporate the *H*-system of navigation into the airborne radar (AN/APS-10)-ground beacon (AN/CPN-6) system described earlier in this chapter. It is necessary to add an accurate ranging unit (a fundamental component of the Shoran interrogator-responsor), and provisions for a split-azimuth indication (one-half of the PPI for each beacon, described in Sec. 17). Sweep expansion, tracking circuits, and ground track determination (by the "pulse-doppler"[1] method) and a pilot's direction indicator greatly improve the accuracy. Such auxiliary equipment can give an *H*-system of precision ±100 ft. without sacrificing the normal radar- and beacon-navigation features. The total weight of airborne equipment is comparable with that for Shoran. No modification of the AN/CPN-6 ground beacons is necessary.

Extensive experiments by RCA have also shown that the instrumental errors of Shoran can be reduced to a value not exceeding ±10 ft.

---

[1] See *Radar Aids to Navigation*, Vol. 2, Chap. 3, Radiation Laboratory Series.

# PART IV

# BEACONS IN THE FIELD

# CHAPTER 20

# INSTALLATION, OPERATION, AND MAINTENANCE

By W. M. Preston and A. Roberts

In this chapter the installation, operation, and maintenance of beacons is discussed. There are four main groups: heavy beacons for fixed ground installations; heavy beacons for shipboard use; airborne beacons; and portable and mobile beacons. Although this classification is not all-inclusive, the principles involved are illustrated sufficiently so that they can be extended readily to other beacon types not specifically included.

## HEAVY GROUND BEACONS

By W. M. Preston

Beacons intended as fixed navigational aids have one primary requirement, reliability. They are frequently called upon to operate continuously; weight and complexity are usually of secondary importance.

**20·1. Choice of Site.** *Altitude.*—It is usually desirable to locate a beacon antenna at the highest point available. Microwave signals, like light, neither follow the curvature of the earth nor bend appreciably around objects that are large in comparison with a wavelength. Therefore, if 360° coverage in azimuth is required of a beacon, it must be located above all neighboring obstacles.

The average maximum range to the radar horizon, as discussed in Sec. 3·4, is given by Eq. (3·9):

$$R = 1.22 \sqrt{h_1} + 1.22 \sqrt{h_2} \qquad \text{nautical miles,} \tag{1}$$

where $h_1$ and $h_2$ are the heights of the antennas of the beacon and interrogator, respectively, measured in feet. If a ground-based beacon is to have maximum range for low-flying aircraft or for ships, then since $h_2$ for the aircraft will be small, it is necessary to locate the beacon antenna as high as possible.

*Interference Nulls.*—Interference between the direct beacon-to-interrogator or interrogator-to-beacon signal and the signal reflected from the earth's surface has been discussed in Secs. 3·1 and 3·2. The resultant periodic fading is nearly always noticeable on an interrogator in an aircraft flying toward a beacon or away from it. The depth of the nulls depends on the reflection coefficient of the earth's surface, which is usually considerably greater for water than for land. There is, therefore, a definite advantage in locating a coastal navigation beacon far enough inland so that the ocean surface is not visible from the beacon antenna. However, unless the coast is very flat, the range may be reduced by the sky line (see below), and the decrease in depth of nulls may not be worth the cost.

The spacing between interference nulls, for a given range and wavelength, is inversely proportional to the product $h_1h_2$. The gain in maximum range by increasing the height of a beacon antenna from 5 to 20 ft is relatively small, but the decrease in separation of nulls by a factor of 4 may be important, as pointed out in Secs. 3·2 and 3·11.

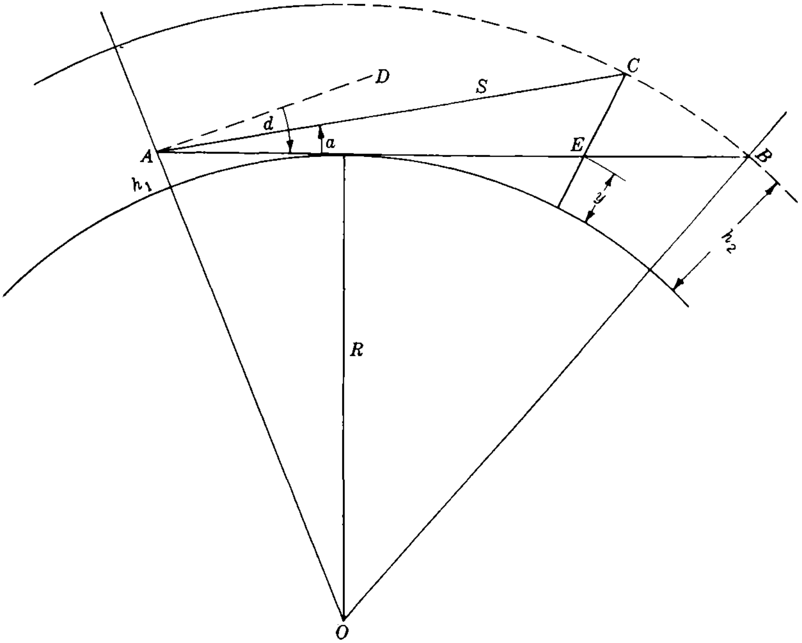


FIG. 20·1.—Reduction in range due to sky-line elevation. $AB$ is the tangent to the sea-level surface of the earth, from the beacon $A$ at height $h$ to the aircraft $B$ at height $h_2$. $S = AC$ is the reduced range when the sky-line elevation angle is $a$ above the sea-level horizon. $O$ is the center of the earth, whose radius is $R$. $AD$ is the horizontal at the beacon.

*Absorption by Trees.*—Little quantitative information is available about the absorption of microwaves by trees, largely because of the impossibility of defining a "tree unit," although the absorption is known to be large. Tests conducted in Florida pine woods indicate little effect on range when a 3-cm beacon is set up in flat country in a very open forest (trees 50 to 100 ft apart). In a dense pine forest, on the other hand, the reduction in range was large, reaching 80 to 90 per cent. This subject is discussed further in Sec. 20·10, below.

*Sky Line: Mapping Beacon Coverage.*—The maximum range as given by Eq. (1) will seldom be attained in practice except over the ocean or exceptionally flat terrain. Hills, trees, or other objects on the sky line reduce the range by intercepting the beam at low angles of elevation. Reduced range may be computed as follows: In Fig. 20·1, let the beacon at $A$ be at height $h_1$. Let $AB$ be a line from the beacon that is tangent to the earth's surface at sea level. The point $B$ is the intersection of this line with a plane that is a distance $h_2$ above the earth's surface. Then $AB$ is the maximum range between a beacon at $A$ and an interrogator at $B$ in the absence of any sky-line obstruction. If some object on the sky line subtends an angle $a$ at the beacon $A$, the range in that direction will be reduced to $AC$. Note that the angle $a$ is measured up from the tangent to the earth, $AB$. The problem is to find $S = AC$ in terms of the heights $h_1$ and $h_2$ and the sky-line angle $a$.

Assuming that the angle $a$ is small,

$$\sin a \approx a \approx \frac{h_2 - y}{S} \qquad \text{and} \qquad S \approx AE. \tag{2}$$

Then from Eq. (1) (with all quantities in the same units)

$$S \approx \sqrt{2Rh_1} + \sqrt{2Ry}.$$

Eliminating $y$ and solving for $S$,

$$S = \sqrt{2Rh_1} - Ra + (R^2a^2 - 2Ra\sqrt{2Rh_1} + 2Rh_2)^{1/2}. \tag{3}$$

Taking $R = \frac{4}{3}$ the mean earth's radius, as is commonly done to allow for refraction, and expressing $h_1$ and $h_2$ in feet and $a$ in degrees,

$$S = 1.22\sqrt{h_1} - 80.0a + (6400a^2 - 194a\sqrt{h_1} + 1.46h_2)^{1/2} \tag{4}$$

nautical miles.

The importance of proper selection of a permanent beacon site justifies a careful sky-line survey from which a coverage diagram can be constructed. To do this, a transit should be mounted at the same height as that proposed for the beacon antenna and close to it. After careful leveling, the transit should be swung around in small steps, taking readings of the elevation angle of the sky line at intervals of a few degrees. The results can be plotted as elevation angle against azimuth. Figure 20·2 shows such a diagram, and Fig. 20·3 the corresponding coverage chart, for a hypothetical site in which the beacon antenna is assumed to have an elevation of 100 ft above sea level, and to be located on the shore with open ocean from 0 to 180° azimuth.

The angles actually measured with a transit will be relative to the horizontal at the beacon, and these are shown on the scale at the left of

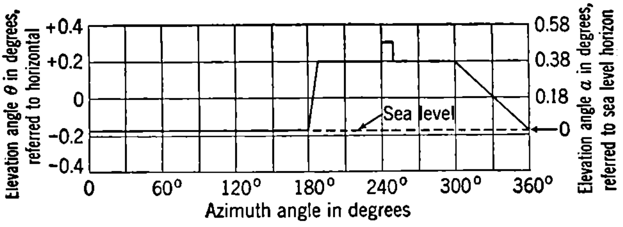


FIG. 20·2.—Sky-line survey. The elevation of the sky line, in tenths of a degree, is plotted against azimuth for an imaginary beacon site. The left-hand scale gives the elevation angle $\theta$ relative to the horizontal at the beacon; the right-hand scale gives the angle $\alpha$ relative to the sea-level horizon, $\alpha = d + \theta$. The beacon antenna height is 100 ft.

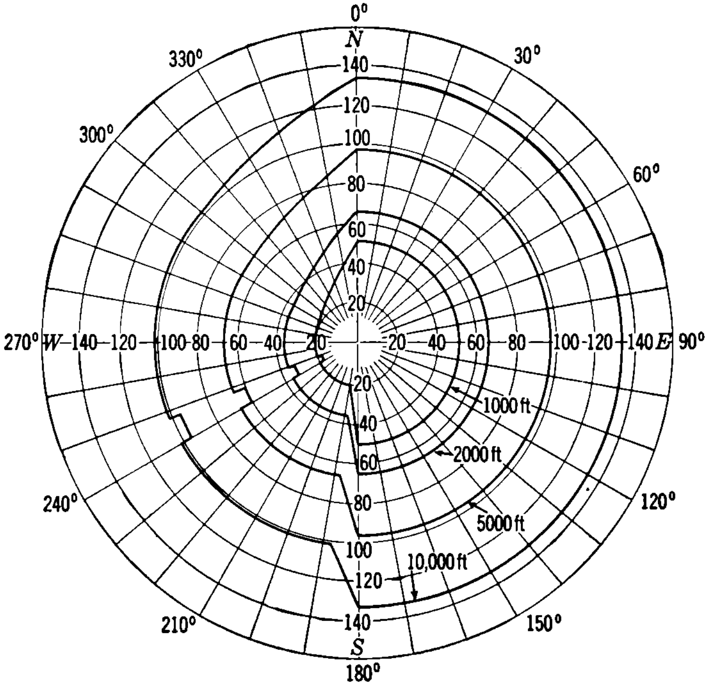


FIG. 20·3.—Beacon coverage diagram, from data from Fig. 20·2. Horizon ranges, in nautical miles, for aircraft at the designated altitudes, are plotted as a function of azimuth angle.

Fig. 20·2. The angles $\alpha$ to be used in Eq. (4) are measured up from the sea-level horizon; they are shown on the right-hand scale. In an inland site where no part of the sky line is at sea level. it will be necessary to compute the angle $d$ (Fig. 20·1) between the horizontal and the sea-level horizon. This can be done by means of the formula

$$d = 0.018\sqrt{h_1} \quad \text{degrees,} \tag{5}$$

where $h_1$, the height of the beacon antenna, is expressed in feet above sea level.

The coverage diagram, Fig. 20·3, was prepared by means of Eq. (4), using the angles $a$ from Fig. 20·2, a beacon height $h_1 = 100$ ft, and interrogator altitudes $h_2$ from 1000 to 10,000 ft as indicated.

*Reflections.*—The discrimination of ground beacons in systems that use interrogation coding may be greatly reduced in poor locations. Consider, for example, an interrogator sending out 1-μsec signals on ordinary search operation, and located at $R$ in Fig. 20·4. The direct signal $RB$ will not by itself trigger the beacon at $B_1$, which we will assume to have a pulse-width discriminator set to accept only signals between 2 and 4 μsec. However, if an object $O$, such as a building, is located at a distance from the beacon such that the time taken by the signal to travel from $R$ to $O$ to $B$ is just 1 μsec longer than the time from $R$ to $B$ directly, the two signals will overlap. Together they will form a pulse 2 μsec long, which will pass through the discriminator. More generally, there will be many objects around a beacon which will reflect or scatter energy toward the antenna. With sufficient amplification the signal received at the beacon will then consist of the direct pulse followed by a long irregular train of weaker, overlapping signals which may continue above the background noise level for 10 or 20 times the length of the direct signal. If the receiver sensitivity is sufficient for triggering by an interrogator at 100 miles, the direct signal will be 40 db stronger when the interrogator is only 1 mile away; reflected pulses have to be more than 40 db weaker than the direct signal if discrimination is to be preserved at this range.

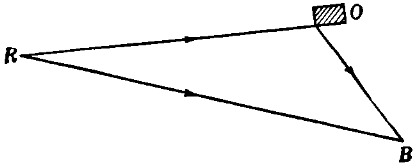


FIG. 20·4.—Echo stretching. Signals from the interrogator $R$ to the beacon $B$, reflected from the object $O$, are delayed relative to the direct signals $RB$.

Shipboard installations are not troubled in this way; the uniform surface of the ocean gives no spurious pulses. Land installations in which the beacon antenna is on top of a high, bare hill are good provided there are no large objects near by. The small vertical beamwidth of beacon antennas is helpful. For example, with a half-power beamwidth of 5° in elevation, objects that lie more than 5° above or below the horizontal at the beacon antenna reflect comparatively little energy into the antenna. The conditions for eliminating spurious interrogating pulses over land are so stringent, however, that they cannot always be met satisfactorily. For this reason, it is important to use receiver circuits that minimize the effect of the secondary signals (see Sec. 8·11).

*Radar Interference.*—In siting a permanent beacon station, an effort should be made to avoid unintentional triggering of the beacon by radar sets in its vicinity. For example, if the station is located near an airport, high-power search or ground-control radar sets or repair shops where airborne radar sets are serviced and tested may be near by. The ground radar sets should be operated on different frequency bands from those to which the beacons respond, but if the radar sets are very close, it may be necessary to install filters in the r-f lines of the beacon receiver.

Alternatively, a trigger pulse from the ground radar can be fed to a suppression circuit in the beacon and the beacon receiver blanked for a short interval during each radar pulse. Repair shops should be located away from the beacon and every effort made to avoid the radiation of strong signals during testing by use of dummy loads or absorbers over the antennas. Methods for suppression are treated in Sec. 16·10.

## INSTALLATION

**20·2. Housing.**—Housing for ground beacons ranges all the way from tents for temporary locations to special buildings for permanent use. When possible, it is advantageous to allow room for a workbench for servicing and for shelves to store spare parts and other supplies near the beacon. The test equipment that is essential for maintaining beacon performance will probably suffice for most repair work; it is, therefore, logical to plan to do all ordinary servicing at the beacon. Provision of adequate heat, light, and space will be repaid by better maintenance.

Beacon stations are comparable to lighthouses. It may even happen sometimes that beacons will be installed at lighthouses; hence, a similar degree of self-sufficiency and a similar approach to the problems of maintenance are indicated.

*Primary Power Supply.*—Heavy ground beacons designed in the United States usually require 60-cps single-phase alternating current at 115 or 230 volts. Stations for intermittent operation can be supplied from motor-driven generators; permanent sites, from central power stations. Although beacons can be designed to work over a reasonable supply-voltage range, their stringent frequency-stability requirements make it wise to provide a power supply free of violent voltage fluctuations resulting from transient loads. If lines must be run a long distance, they must have sufficient capacity to prevent undue voltage drop at the maximum load imposed by the beacon.

*Arrangement of Associated Test Equipment.*—The guiding principle in the arrangement of external test equipment in a permanent beacon station should be convenience. Test equipment not built into the beacon can be located on shelves or brackets—as shown, for example, in Fig. 15·11, Sec. 15·12. Every control to be used for adjusting the reading of a meter must be located so that the meter can be watched as the adjustment is made.

**20·3. Antenna Installation.**—Once a site has been chosen and housing has been provided for the beacon, the installation of the antenna must be carefully planned.

*Mast.*—Often the antenna must be mounted on a mast in order to clear its immediate surroundings. The mast must have adequate strength and be strongly guyed so as to withstand the maximum wind velocities expected, with a suitable safety factor. In high latitudes, further allowance must be made for ice loading.

*Deicing.*—Although ice is relatively transparent to microwaves, water absorbs them strongly. A thin layer of ice may absorb rather little power, but it may reflect enough to produce serious long-line pulling of the transmitter frequency. If no provision for heating or other deicing is provided the antenna must be readily accessible for manual cleaning in regions where snow and ice are encountered.

*R-f Lines.*—It is always advantageous to keep the r-f lines that feed the antenna as short as possible. The maximum permissible length from the equipment to the antenna is determined by two factors: the power attenuation and the danger of long-line effects. The latter depend on the antenna match, the line length, and the transmitter-tube characteristics. A maximum allowable line length, which takes into account both attenuation and long-line effects, should be specified by the manufacturers for each type of beacon.

*Pressurizing.*—The need for pressurizing air-filled transmission lines has been set forth in Sec. 7·7. In field installations, the r-f transmission lines and the antennas should be made a common system for gas pressure. A pressure gauge at the beacon should be provided, and either a convenient inlet valve for a hand-operated pump, or a permanently connected motor-driven pump. In order to prevent condensation of moisture only dry gas should be introduced into the lines. Air dried by passage through silica gel or other suitable desiccators is thoroughly satisfactory. In a permanent installation, automatic pressure maintenance is advisable. This is achieved by a pressure-operated switch which is connected to the antenna lines and controls a motor-driven pump so as to keep the r-f lines filled with dry air between specified pressure limits.

*Multiple Installations and Crystal Damage.*—In many cases, a station will consist of a number of beacons on different frequencies, with, possibly, dual installations on each frequency. When uninterrupted operation is essential, a dual installation provides a complete stand-by unit which can be switched in rapidly. In the resulting clutter of antennas, care must be taken to avoid damaging the receiving crystal in one beacon by the transmitter of another. Beacons using waveguide will ordinarily be protected against frequencies substantially lower by the cutoff property of waveguide; there is, however, a possibility that higher harmonics may be present with sufficient power to cause trouble. If there is any doubt, a direct calculation of the power received should be made,[1] as follows:

$$P'_R = \frac{\lambda^2}{(4\pi)^2}\frac{P_T G_T G'_R}{R^2}, \tag{6}$$

where the notation is that of Chap. 2. It is assumed that the two antennas are in the same horizontal plane, so that the receiver is in the direction of maximum gain of the transmitter.

In a typical case, $\lambda = 10$ cm $= 0.33$ ft; $P_T = 10{,}000$ watts, $G_T = G'_R = 10$. For $R = 25$ ft, from Eq. (6) the received power $P'_R = 1.1$ watt, which is just about the safe upper limit for many crystals. The antennas, therefore, must be separated by at least 25 ft; if they must be closer, they should be staggered in height, so that the receiver of each lies far enough below or above the transmitter of the other to be out of the main lobe of the pattern.

*Shielding by Masts.*—With multiple-antenna installations, care must be taken that the different antennas or their masts do not shade each other. In Fig. 20·5, $A$ is a beacon antenna; $B$ is a cylindrical pole, of metal or other material opaque to the beacon radiation. The diameter of $B$ is $S$, and its distance from $A$ is $a$. At a distance $b$ from the pole, the latter will interfere with the antenna pattern over a region $R$, which is at least somewhat larger than the geometrical shadow.

Let us assume that $b \gg a$, $a \gg \lambda$, that $A$ can be regarded as an infinite line source of radiation, and that the pole $B$ is a perfect absorber for microwaves. Then it can be shown by means of diffraction theory[2]

---

[1] The inverse square law may be applied for distances from the antenna greater than $2d^2/\lambda$, where $d$ is the antenna aperture. For linear arrays, Eq. (7·1), Sec. 7·4, gives, for the minimum distance at which the inverse square law holds, the value $G^2\lambda/2$. At shorter distances the inverse square law breaks down, the actual received power being less than that predicted.

[2] See, e.g., Jenkins and White, *Fundamentals of Physical Optics*, McGraw-Hill, New York, 1937, p. 199.

that the intensity will nowhere be down by more than 3.5 db if

$$S < 0.35 \sqrt{\lambda a}, \tag{7}$$

where $\lambda$ is the wavelength and $S$, $\lambda$, and $a$ are all measured in the same units of length. This gives for the case $\lambda = 3.3$ cm and $a = 3$ meters, $S = 0.35 \sqrt{3.2 \times 300} = 11$ cm. At a large distance behind a pole of this diameter, therefore, the intensity will nowhere be more than 3.5 db below what it would be if the pole were absent.

Equation (7) can be considered only a rough guide when the assumptions are not valid. In particular, if the pole $B$ is of metal and relatively close to the antenna, resonant currents may be set up on it which will reradiate energy and modify the antenna pattern considerably, not only behind the pole but in all directions. It should, however, be possible to eliminate resonant currents by coating metallic surfaces with a material that absorbs microwaves.[1]

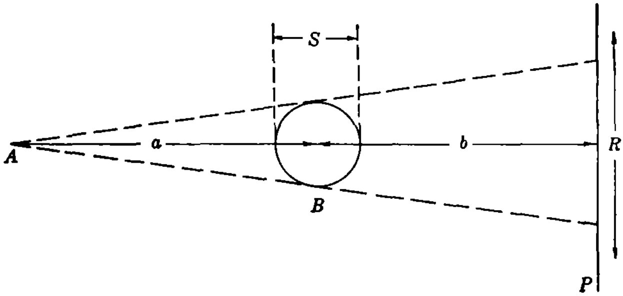


FIG. 20·5.—Antenna shielding by neighboring objects. $A$ is a linear-array beacon antenna, axis perpendicular to the plane of the diagram. $B$ is a cylindrical pipe of diameter $S$ and axis parallel to $A$. $R$ is the region of the diffraction pattern observed on a plane $P$.

## 20·4. Operational Check.

—Following the completion of a new beacon installation, particularly one intended as a permanent navigational aid, it may be desirable to make an operational test of its performance. For an airborne-interrogator–ground-beacon system, one or more flights can be made to check extreme range in various directions and at various altitudes, minimum range, and decoding. In order to get useful quantitative information, however, a flight test of an interrogator against a beacon must be made with great care. The characteristics of the interrogator and the beacon, such as power output and receiver sensitivity, must be known accurately. Otherwise, a test with an interrogator of above-average performance will overestimate the beacon performance. In addition, temporary conditions of anomalous propagation may result

---

[1] See *Principles of Microwave Circuits*, Vol. 8, Chap. 11, Radiation Laboratory Series.

either in abnormally large ranges at low altitudes, or sometimes in abnormally low ranges.

Because microwave propagation under normal conditions is so nearly rectilinear, the coverage of a beacon can be predicted accurately by the ground-survey method discussed in Sec. 20·1. All other characteristics of the beacon can be measured accurately by suitable test equipment at the beacon; if these meet specifications, the beacon range can be calculated for any type of radar. In the absence of precise data on the propagation characteristics to be expected during a test, the laws of optics are a better guide to the range to be expected than a flight test made at random. For these reasons, operational flight tests seldom serve a useful purpose.

## OPERATION AND MAINTENANCE

**20·5. Local Performance Checks.** *Personnel.*—The use of heavy ground beacons as continuously operating navigational aids for radar-equipped aircraft has encouraged their design as automatic devices requiring no operator. Since weight is ordinarily secondary to reliability, recycling relays are usually provided to reestablish operation following a transient overload or power failure. Some installations include a complete stand-by beacon, and provision may be made for switching over to it automatically if the main unit should fail. In this way, the need for constant attendance can be eliminated.

However, continuous operation requires routine performance checks at stated intervals, as well as regular preventive maintenance. Since, in addition, it is usually desirable to make all ordinary repairs at the beacon site, a highly skilled maintenance man should be available at each beacon installation. This contrasts, for example, with the problem of maintaining small beacons in aircraft, in which case major equipment troubles are frequently eliminated by replacement of the whole unit, which is then sent to a central depot for servicing by specialized personnel.

*Over-all Check of Performance.*—A complete over-all check of beacon performance indicates whether or not the equipment puts out an r-f signal (1) of power greater than the minimum specified; (2) at a frequency within the specified limits; (3) with the proper number and spacing of code pips, when it is triggered by an r-f signal (4) of the minimum specified power; (5) of frequency within the receiver band; (6) with pulse coding (if coding is provided) only within specified limits, and (7) between the lower and upper limits of pulse-repetition frequency. A satisfactory test of this nature shows that the beacon is operating properly; failure at any point indicates the need of more detailed tests to localize the trouble.

A considerable amount of test equipment is necessary to conduct such an over-all check and it must be of high quality and good design since absolute rather than relative values of quantities are required and some of the tolerances are close. Finally, the simpler and more automatic we try to make the process of measurement, the more complex the design of the test equipment becomes. As a result, it may be advisable, after a study of the probabilities of various failures in a beacon, to devise a simpler and less inclusive over-all check which can be carried out at more frequent intervals.

*List of Test Equipment.*—The list of test equipment which follows includes the items needed for complete maintenance of a typical microwave ground beacon:[1]

1. Test set or pulsed-signal generator.
2. Synchroscope.
3. Transmitter frequency standard.
4. Volt-ohmmeter.
5. Voltage divider.
6. Dummy load.
7. Local-oscillator frequency standard (when superheterodyne receiver is used).
8. Standard tube tester.
9. Fluxmeter (for checking the field strength of the magnetron magnet).
10. Aural monitor (with provision for switching to any one of a number of test points).

Of these items, the first five are the most essential. The dummy load is used rarely, but it is needed when antenna line mismatch is suspected and when it is undesirable to radiate power during testing of the transmitter. The local-oscillator standard cavity will be part of the receiver if the latter has local-oscillator AFC; it will be unnecessary if the frequency meter in the test set is accurate enough to check the local oscillator. A standard tube tester saves time and helps to anticipate tube failure, but its provision may not be justified except in a large beacon station. Magnet failure is unusual, but difficult to discover without a fluxmeter. An aural monitor may be classed as a convenience; it is useful both for testing and for continuous monitoring.

*Over-all Check.*—Using the test equipment listed above, the complete over-all check may be performed as follows:

1. Trigger the test set from the synchroscope. Adjust the output of the test set to a frequency within the receiver band, to a pulse code

---

[1] The items themselves are described functionally in Secs. 15·12 to 15·20. Although specialized units, such as the test set, may be built into the beacon cabinets instead of being packaged separately, their functions will be essentially the same in either case.

that should trigger the beacon, and to the power level corresponding to the minimum specified receiver sensitivity. Connect the voltage pulse on the magnetron, from the appropriate test point, to the vertical deflection plates of the synchroscope. If a pulse is visible, it indicates that the beacon is being triggered by the minimum specified power.

2. Vary the frequency of the input signal over the whole specified receiver band. The beacon should be triggered continuously.
3. Vary the coding of the input signal. The beacon should start to trigger at the lower limit of the pulse-coding tolerance and cease to trigger above the upper limit. (The synchroscope may be used to monitor the pulse code, and the aural monitor to indicate when triggering commences.)
4. On a suitable synchroscope sweep speed, check the number, width, and spacing of the beacon code pips at the modulator-voltage-pulse test point.
5. Trigger the beacon from the synchroscope at some convenient point, such as the coder input. This is done to release the test set so that its r-f output can be connected to the r-f test point on the transmitter line. The power monitor in the test set is now used to measure the beacon transmitter average power. From the repetition rate and duration of the voltage pulse of the modulator, the pulse power can be computed.
6. Check the transmitter frequency by tuning it to peak the output of the standard cavity. Then vary the triggering repetition rate over the specified limits. The output frequency should remain within the required limits. *Relative* frequency changes can be measured with sufficient accuracy with the frequency meter in the test set, still connected as in the preceding paragraph.

Mention should be made here of a difficulty in making a satisfactory over-all check which was not mentioned in the foregoing description. It is obviously desirable, in such a check, to introduce the triggering signal at the receiving antenna and to check the transmitted power at the transmitting antenna. Only in this way will the test include the condition of the antennas and r-f transmission lines. To do this requires (1) a pickup dipole or other device located close to each beacon antenna, adjusted to a known coupling, and (2) r-f lines from each of the pickup elements (in the case of systems with separate receiving and transmitting antennas) down to the equipment. When the antenna lines are long, about 50 ft, the test lines themselves introduce an uncertainty at least equal to that of the beacon lines proper. For such distances, the attenuation and general unreliability of coaxial line is generally too great to be tolerated at 10 cm and even more so at shorter wavelengths, and four lines of waveguide running up to the antenna make a heavy and clumsy installation. Recently, duplexers of bandwidth sufficient for microwave beacon use have become available; these switches make it possible to have a single antenna line and a single test dipole line.

The alternative practice has been to measure receiver sensitivity and power output at test points close to the beacon equipment. Waveguide losses are brought about by physical injury, which can be spotted by careful inspection, or by dirt or water inside the guide. If the antenna and guide lines down to the equipment are sealed off and maintained at a pressure above atmospheric, no increases in attenuation will occur. Operating practice has been to keep both antenna and lines pressurized and to assume that they are in good condition.

Another method of measuring the total line losses has been used with some success but is applicable only to systems having separate transmitting and receiving antennas. A bar or other device is clamped temporarily in a reproducible position where it increases the coupling between the two antennas to a figure of 20 to 30 db. The receiving line is disconnected at the beacon and connected to the test set, which is then used to measure the average received power when the transmitter is running. This is compared with the power from a directional coupler in the transmitting line near the beacon. The difference, corrected for the coupling between the antennas and the coupling of the directional coupler, is the sum of the losses in both antenna lines. If the coupling between the antenna lines is not known, changes in line loss from day to day can still be detected.

*Maintenance Schedule.*—It is impossible to draw up a maintenance schedule to apply to heavy beacons in general since it must be based on a study of the performance and reliability of a particular type of equipment. It should probably call for a daily over-all check of the type outlined above, with whatever additional checks experience might dictate. At longer intervals, a thorough overhaul of all components will serve to anticipate trouble. For example, when data on tube life are available, it may be advisable to schedule a change of tubes shortly before the end of their average useful life.

**20·6. Remote Performance Checks.** *Remote Monitors.*—Ground beacons will frequently be located at or near airports, Coast Guard stations, or similar establishments. The best location for the beacon may be some distance from the base—on a neighboring hill, for example. Since beacon equipment is automatic and requires attention only when something goes wrong, a remote monitor at the base will save time by making it possible to check operations without going to the actual beacon site.

In most respects, the design of a remote monitor is straightforward and depends primarily on the complexity and expense which appear to be justified. The simplest monitoring can be done by means of wire connections between the beacon site and the base. For example, a remote indicator light can be used to give warning when the high voltage has gone off because of power failure, overload, or other cause. A more complicated system might contain provision for turning on the test set described above, preadjusted to feed in an r-f signal to the receiver at a power level approximating that corresponding to minimum receiver sensitivity. An aural monitor, consisting of amplifier and loudspeaker, can serve as indicator. Connected to the output of the transmitter frequency standard, it would show whether or not the beacon was putting out a signal on the correct frequency, and give some idea of the relative power output. This would be a simplified and reasonably satisfactory over-all check, but it would need to be supplemented by periodic and more extensive checks at the beacon site.

Another monitoring system involves a calibrated interrogator at the base. This requires first of all a clear line of sight from the monitor antenna to the beacon antenna. There are, however, other limitations. If the beacon has a high-gain antenna and is located on a hill considerably higher than the monitoring station, the latter may be below the main lobe of the beacon antenna pattern. As a result, an impractical amount of power may be required to trigger the beacon. In addition, if the path between monitor and beacon is over water, interference between the direct and reflected rays (Sec. 3·1) will cause large variations in signal strength due to tidal changes in the level or varying surface conditions of the water. While variations over land will be much less because of the lower reflection coefficient, they may still limit the accuracy of the system.

This type of monitor may be built primarily from radar components. It is probably simpler to use separate receiving and transmitting antennas; these can be relatively small paraboloids mounted so as to point at the beacon. The received signal is led to the receiver through a calibrated variable r-f attenuator. The indicator may be a synchroscope used for type A presentation. The synchroscope is also used to trigger the modulator. A second calibrated variable attenuator in the transmitter line makes it possible to vary the r-f output. With directional-coupler test points in the receiver and transmitter lines, a test set of the type described in Sec. 15·13 will make it possible to calibrate the power output and receiver sensitivity and to measure the frequency of the monitor transmitter and of the beacon transmitter. By decreasing the monitor power output until the beacon ceases to trigger, the beacon receiver sensitivity can be checked. The bandwidth of the beacon receiver can be tested if the monitor has a tunable transmitting tube. The power output of the beacon can be checked by varying the attenuator in the receiving line of the monitor until signals reach a calibrated amplitude. Interrogation coding may be checked by providing a suitable coder for the monitor modulator, giving pulse codes than can be varied above and below the specified pulse-code limits.

*Quantitative Flight Checks.*—It has been said above that routine flight tests of ground beacons, using standard airborne radar equipment, are of doubtful utility. Beacon performance can be checked more easily and accurately by using test equipment. However, occasions will arise—for example, in the design of new beacons—when flight tests will be desirable to verify the calculated performance.

By modifying a standard airborne interrogator along the following lines, a satisfactory airborne testing device, which can be monitored and controlled while in flight, may be built.

1. Insert a power divider between the transmitter tube and the TR switch. Couple a thermistor bridge or other power-measuring device into the transmitter line by means of a directional coupler or similar device.
2. If the frequency is to be varied, a tunable transmitter tube is required. Its frequency can be measured by a frequency meter in the line between the directional coupler and the power monitor.
3. Mount a pulsed-signal generator of the type described in Sec. 15·13 near the interrogator, and trigger it from the synchronizer. Connect the r-f output of the signal generator to the antenna line through a second directional coupler so as to send signals toward the receiver.
4. Add an A-scope indicator, if the interrogator does not have one already.
5. Replace the antenna by a smaller reflector giving a half-power beamwidth of at least 20°. Because of the motion of the aircraft, it is difficult to make quantitative measurements in flight if the antenna beamwidth is small.

In operation, beacon-receiver sensitivity can be tested by reducing the radar power output and determining the level at which the beacon ceases to reply. Bandwidth can be checked by varying the radar transmitter frequency. Beacon power output can be determined by matching in amplitude the beacon signal on the A-scope with an artifical signal from the signal generator.

It is well to work over land, rather than over water, and to fly at altitudes and ranges well above the radar horizon, in order to reduce variations of signal strength due to interference (Sec. 3·1). Range measurements made with reduced radar power output and sensitivity can be extrapolated to give reliable range by means of Eqs. (2·2) and (2·3), Sec. 2·5.

The precision of such measurements is reasonably good. An idea of the variations to be expected can be obtained by an examination of the dispersion of the experimental points in Fig. 2·4 about the inverse square line to be expected from theory. The mean of a series of careful measurements made on a flight should predict the system performance with a probable error not exceeding 2 db.

## SHIP BEACONS

BY W. M. PRESTON

Since weight and complexity are usually secondary to serviceability and reliability, shipborne beacons of the type intended as navigational aids are in many respects similar to the heavy ground beacons discussed above. Here attention will be called only to the principal differences between installations on shipboard and those on land.

**20·7. Installation, Operation, and Maintenance.** *Location of Equipment.*—As compared with ground installations, space for beacons on shipboard is limited. By mounting components in drawers that pull out and lock and by making it possible to get at all components from the front of the cabinets, however, most beacon maintenance can be done right at the equipment. The equipment can be mounted close against a bulkhead out of the way, and lateral shock-mount brackets added where necessary.

Since the antenna is usually mounted as high on the mast as possible, the antenna lines are long at best. Moreover, installation of the antenna lines is frequently difficult, particularly if they must be run through one or more decks. It is, therefore, wise to locate the equipment on an upper deck, if space can be found, in order to keep the length of the antenna lines to a minimum.

*Antenna and Antenna Lines.*—If all-round coverage is required, the antenna must be mounted near the top of the mast. Where this is impossible, it may be put on a yardarm. Equation (7) may be used as a guide in determining whether or not the mast or other vertical cylindrical structures will cause an appreciable shadow. If other radar equipment is near by, consideration must be given to the dangers of damaging the crystal of the beacon receiver and of picking up excessive interference. It is usually possible to find a satisfactory solution because of the directivity of the antennas, as long as they are not located on the same level.

The range of a beacon on shipboard will be given by the simple formula, Eq. (1), since the horizon is sea level, except as the antenna pattern may be obstructed by masts or other elements on the ship itself. Interference nulls will always be a prominent feature because of the high reflection coefficient of sea water (Sec. 3·2). Pulse stretching at sea will ordinarily be negligible.

For wavelengths of 10 cm or less, the r-f lines will normally be waveguide, which has lower attenuation and which is far more rugged than coaxial line. It has been common practice to solder all joints or to seal them with gaskets and to maintain a positive pressure of dry air in the lines at all times. This prevents internal corrosion and the losses that result from condensed moisture and it also gives an indication of any break in the lines. Most installations will require specially fabricated bends. Because the bending of waveguide requires equipment and skill, it must be done by experienced personnel.

*Reflections.*—When ships are isolated, reflections give very little trouble at sea. There are, however, circumstances under which reflections may cause difficulty with shipboard beacon installations.

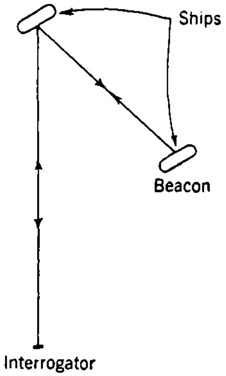


FIG. 20·6.—The origin of spurious interrogations and responses of shipboard beacons by reflections from another ship. A beacon on one ship may be interrogated by a signal reflected from another, and its response may in turn be reflected from it.

When the ship carrying the beacon is close to a coast line, echoes may give rise to disturbances, just as in the case of ground installations. This, however, is a relatively trivial matter.

The large radar cross section of ships, however, may cause confusing reflections when ships are within only several miles of each other. Difficulties may arise through reflection of either the interrogation or the response, or both.

In Fig. 20·6, suppose Ship $A$ carries a beacon, and Ship $B$ is a few miles away. An interrogation reflected from $B$ may trip the beacon on $A$. The beacon response is then observed by the interrogator at an incorrect range. It is possible also that a direct signal from the interrogator may trip the beacon, and the beacon response then may be reflected from $B$ to the interrogator giving incorrect range and bearing. Finally, both the interrogation and response may be reflected from $B$.

The seriousness of such spurious indications depends upon the angular discrimination of the interrogator and the geometry of the situation. The spurious indication will always be at a range greater than the correct range, so that a careful operator can distinguish between them. Angular discrimination alone is of no assistance when both interrogation and response are reflected. When many ships are assembled, as in naval formations, spurious "beacons" may, because of reflections, appear on every ship in the formation at short ranges. Relative intensity and range are the parameters by which the interrogator operator can distinguish true response from reflections. In practice, it is not often difficult to distinguish true from false indications.

Response coding is of great assistance in distinguishing beacon replies from radar reflections that may break through when the interrogator is adjusted for beacon reception. For example, suppose an airborne or shipborne radar used as a beacon interrogator has a maximum free-space range of 60 miles on a large ship. At 6 miles the radar response is then 40 db above the minimum useful response. If the selectivity of the interrogator-receiver, or channel rejection, between radar and beacon signals is 40 db, the radar reflection from the ship carrying the beacon will begin to appear on the screen at 6 miles even though the interrogator is tuned to beacon signals. If the beacon replies are uncoded this may result in serious confusion; response coding usually will eliminate any possibility of such confusion.

Distortion or garbling of interrogation coding by reflections is possible in ship formations; but it is relatively infrequent, since the geometrical conditions for this occurrence are stringent. On very large ships, pulse-width coding may give trouble, since echoes from distant portions of the ship may act to stretch incoming pulses.

Time-varied gain of the interrogator-receiver will eliminate or greatly reduce spurious effects due to reflections. The suppression of spurious pulses in the beacon receiver is of less value than with ground receivers, since the interfering echoes come relatively long after the direct signal.

*Interference.*—A thorough search for possible interference should be made after installation of a beacon on a ship. The type of interference to be looked for includes not only unwanted triggering of the beacon by other radar or communication equipment, but interference by the beacon with other radar and communication sets. It has been demonstrated that it is practical to eliminate virtually all interference that does not actually enter the antenna by proper design, involving good grounds and shielding and filtering of power lines.

It has been the rule to set the beacon transmitter frequency in a clear channel in order to avoid interference with radar sets, and to operate ship radars on different bands from airborne equipment so that a beacon intended for airborne interrogators should not be triggered by sets on the ship. If, however, one or more ship radars cause interference which triggers the beacon through the r-f system and which cannot be eliminated by filters, it will be necessary to use a suppression circuit (Sec. 16·10).

*Operation and Maintenance.*—The operation and maintenance of shipboard beacons are similar to those of heavy ground beacons. It is usually possible to mount close to the beacon all test equipment necessary to make a complete over-all check of operation similar to that described in Sec. 20·5. The same adherence to a definite preventive maintenance schedule is required if continuous, reliable operation is desired. The antenna and lines on shipboard are particularly liable to damage by strain and vibration, and they are seldom readily accessible. It is, therefore, desirable to be able to measure the over-all line and antenna losses by means of a separate waveguide connected to a pickup probe at the antenna, as described in Sec. 20·5.

## AIRBORNE BEACONS

BY W. M. PRESTON

Airborne beacons are generally characterized by their light weight and, in comparison with ground or ship installations, by their simplicity. They are operated intermittently, although quite possibly for many hours at a stretch; during operation little or no maintenance is possible.

**20·8. Installation.** *Location of Antenna.*—Every type of aircraft presents a new problem in antenna installation, and the solution usually involves compromises. The stringent requirements of minimum aerodynamic drag must be met, and at the same time a location is required which gives the best possible coverage and a mounting that is structurally practicable. The lowest point of the underside of the fuselage, in normal flight aspect, is frequently the obvious choice. This location may be ruled out if it is shielded by external wing tanks or radomes, and it may be relatively poor in low-wing models because the wing will obscure the antenna when the aircraft banks.

Trials of double installations have been made, for example, with two receiving antennas located on opposite wing tips and connected to the receiver in parallel, and two transmitting antennas similarly located. Although it is possible thus to insure that coverage is obtained in all directions, the interference pattern in directions from which both antennas are simultaneously visible is complicated by many lobes. These lobes are probably intolerable for automatic-tracking radars like the SCR-584 which require continuous information. Such antenna installations may, however, be used with separate receivers for each antenna, with mixing of the video output signals.

The mocking-up of an antenna installation in a new aircraft type is a major undertaking, and it must be done and the testing carried out in a well-equipped laboratory. The approved mockup must be followed carefully in later installations, since small changes in location may affect the pattern. The difficulty increases with decreasing wavelength of the system.

*Location of Equipment.*—Aircraft beacons are usually designed with small control boxes that can be mounted within convenient reach of the operator. The equipment proper may then be located wherever space is available, subject to the stipulation that the lengths of the antenna lines be held to a specified minimum.

*Reflections.*—Airborne beacons may be triggered by reflections of the interrogating signals from the ground, or, more infrequently, from other aircraft. An interrogator in the same aircraft can trigger an airborne beacon, even if there is a direct suppression connection, if the signal reflected from the ground returns with sufficient intensity after the suppression period is over. Suitable choice of the suppression interval, so that the reflected signal is too weak to trigger the beacon at altitudes for which the reflection time exceeds the suppression interval, is necessary.

With ground interrogators, reflected interrogations from large objects on the ground may reach the airborne beacon after the direct signal. This may result in apparent double pulses being received at the beacon; if double-pulse coding is used, codes may be simulated, so that the beacon may be triggered by interrogators that are not meant to trigger it. If the decoder is not suitably designed, the reflected pulse, appearing between the two direct pulses of the code, may prevent a response.

Beacon responses to an airborne or ground interrogator can be reflected from the ground and thus appear at incorrect ranges and azimuths. As in other cases, angular discrimination in the interrogator and time-varied gain of its receiver will tend to remove these interfering signals.

*Interference.*—In locating the beacon antenna in the mockup of a new-type aircraft, care must be taken that sufficient power from other radar equipment to damage the receiving crystal does not enter the beacon. This is unlikely to occur unless a radar antenna points almost directly at the beacon antenna. If an actual measurement of the power reaching the beacon crystal shows it to be excessive, either an r-f filter must be installed or the beacon antenna must be relocated. Triggering of the beacon by other radar equipment can be eliminated by using a suppressor blanking pulse.

**20·9. Operation and Maintenance.** *Personnel Requirements.*—The operation of an airborne beacon is usually extremely simple and may consist of little beyond turning it on or off. In any case, its operation will be a minor duty of a pilot, radio operator, or other crew member who cannot be expected to be highly trained in maintenance.

Actual maintenance will be done on the ground. Since it is difficult to make extensive repairs of radar equipment inside the aircraft, it is desirable to provide for making a very simple over-all check of beacon operation at the aircraft. If any major difficulty is encountered, either the troublesome component or the whole beacon should be replaced immediately. This check can be made by relatively unspecialized personnel and the defective components sent to a depot to be serviced by capable maintenance men under favorable working conditions.

*Built-in Monitoring Equipment.*—For airborne beacons, built-in monitoring equipment is usually held to a minimum by limitations on space and weight and may be entirely absent. Perhaps the most useful single test feature that may be supplied is a push-button-operated video triggering circuit to actuate the equipment beyond the decoder and a tuning control and meter indicator for setting the transmitter frequency.

*Preflight-check Test Equipment.*—If the maintenance organization suggested above is followed, relatively simple test equipment for checking a beacon in an aircraft is wanted. The quantities to be checked are (1) power output, (2) frequency, (3) response code, (4) receiver sensitivity, (5) receiver band coverage, and (6) interrogation coding. Moreover, because of the considerable chance of injury to the antennas due to their exposed position, or to the antenna lines due to the extreme vibration to which aircraft are subject, an over-all check should certainly include the antennas and their lines.

These specifications will almost certainly require a special piece or set of test equipment designed for this purpose alone. It must fulfill most of the major functions of the test set and synchroscope described in Sec. 20·5, such as providing a pulsed signal of known frequency and variable calibrated power output, measuring the frequency and power o the beacon signal, and examining the code. Ideally, it would be battery-operated; alternatively, it may be run from mobile engine-driven generators of the type expected to be available at airports. It must either be readily portable, or it must be carried by a test truck, and the latter must be available in the required locations. A reproducible method of coupling the test set to the beacon antenna is required. This problem has been solved in one case by a "black," or completely absorbing, "hat" to be fitted over the beacon antenna, with pickup dipoles built into it.

*Bench Test Equipment.*—Complete test equipment must be supplied for general maintenance work. If the procedure suggested is followed, servicing will be done in a repair shop under favorable conditions. Hence, flexibility and reliability will be required of the test equipment and weight will be secondary. The same set of test equipment will be used to service many beacons and possibly other radar gear. It will be essentially that listed in Sec. 20·5.

*Remote Monitors.*—Although in most cases the system of preflight checking described above will be entirely sufficient, occasions may arise when it is desirable to obtain an operational check of an airborne beacon just before takeoff, or in flight, while the aircraft is still near its base. A monitoring station for this purpose can be constructed from a modified interrogator. It must have a rotatable or omnidirectional antenna in order to follow the aircraft easily. The extent of the modification required depends, of course, on the completeness of the over-all check desired. A power divider in the transmitter line, a tunable transmitter, and an attenuator in the receiver line, together with a test set for calibration, will make possible rough quantitative checks on beacon power output and receiver sensitivity. It should be pointed out that a unit of this type may be complicated and expensive.

## PORTABLE AND MOBILE BEACONS

BY A. ROBERTS

**20·10. Installation, Operation, and Maintenance.**—The installation of a portable or mobile beacon differs from that of a large ground beacon

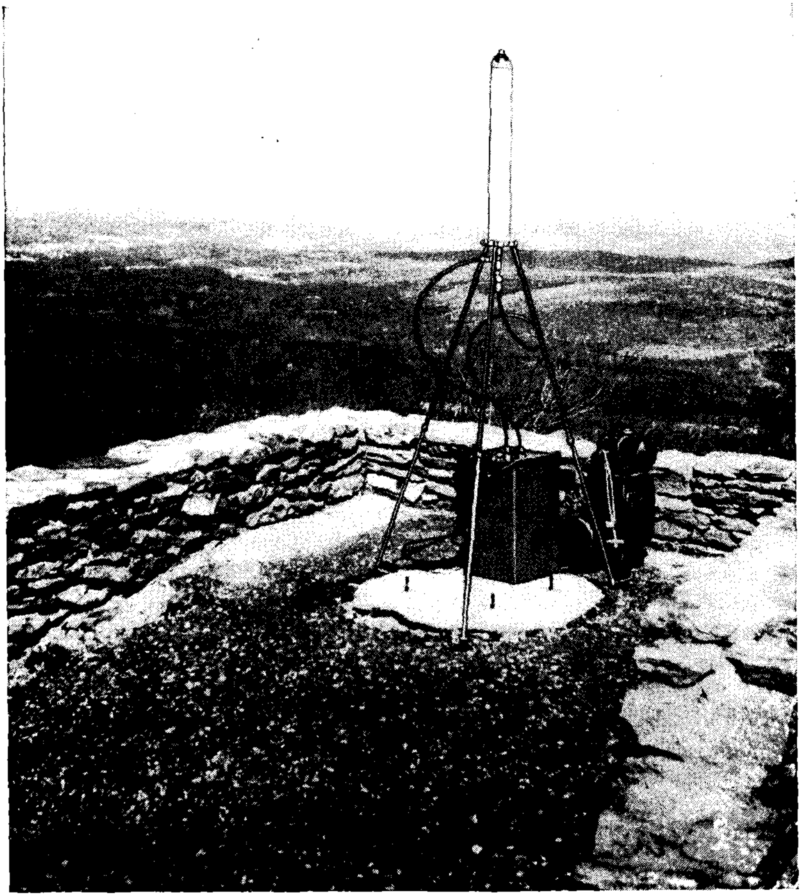


FIG. 20·7.—An ideal portable beacon location: the top of Mt. Wachusett, Mass.

in that it is likely to be temporary rather than permanent. There may, therefore, be less time to survey available sites so that the best one may be chosen. In a temporary installation, furthermore, the erection of high masts may not be practical, unless these are of the sectional type and provided as part of the beacon equipment.

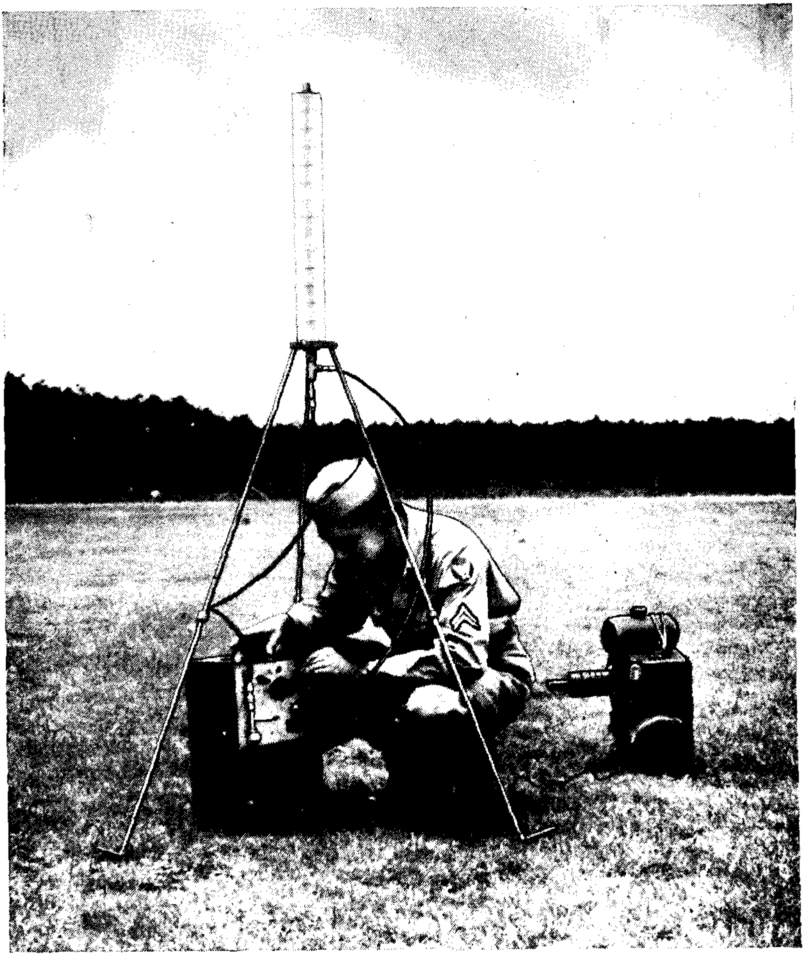


FIG. 20·8.—A very good portable beacon location. The sky-line elevation is raised somewhat by the trees.

The considerations governing the choice of site are, of course, identical with those for permanent installations. In view of the possible necessity of setting up a beacon hurriedly in an unsurveyed location, it is important

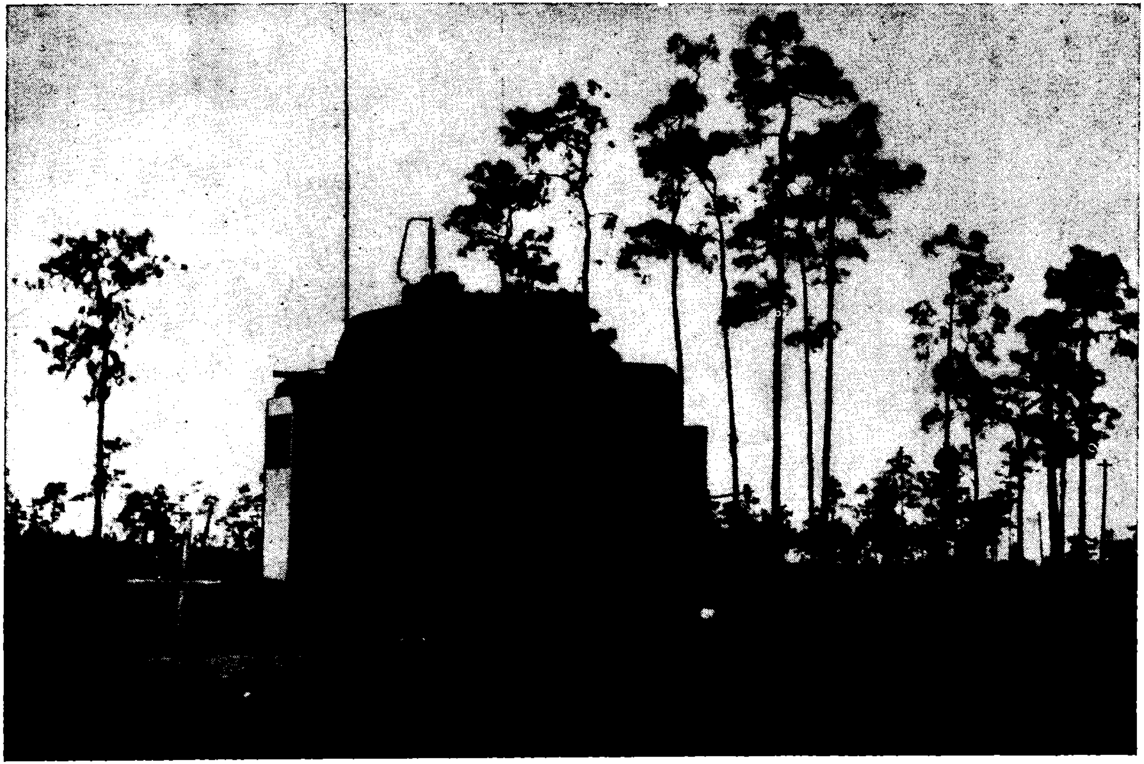


FIG. 20·9.—A very good beacon location. The trees are scattered and the terrain flat.

for the operator to be well acquainted with the requirements for satisfactory beacon sites and with the performance of his equipment under conditions of unfavorable location.

The height of the beacon antenna in temporary installations is generally determined by the antenna support supplied with the equipment. In rolling or hilly country the operator should, of course, try to select a site on top of a hill rather than in a valley.

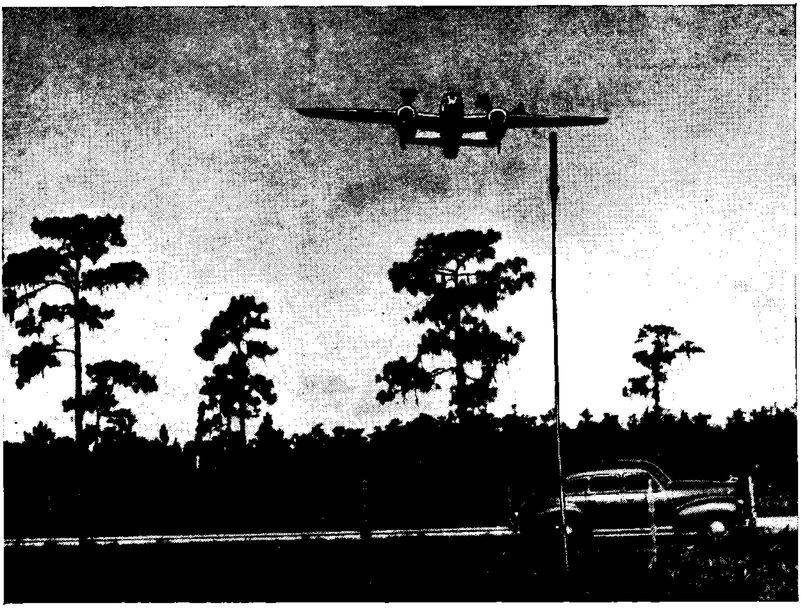


FIG. 20·10.—An excellent beacon location. The antenna mast is nearly as high as the distant trees. Homing on a portable beacon may be very accurate.

The effect of interference nulls may be very noticeable for portable beacons, even if the reflections occur on land rather than on water. The possible height of the beacon antenna is not great and for this reason the nulls will be relatively far apart. Location in flat, open country favors interference nulls; location on a hilltop suppresses them.

The effect of trees on range must be well understood by the operator, and this can be learned only from experience. Figures 20·7 to 20·17 show portable-beacon locations of varying excellence and give an idea of the effect of trees on reducing range. The reduction of range by trees may result either from attenuation or from an elevated sky line. At the lower frequencies, polarization is of importance. Horizontally polarized radiation will be attenuated less by trees than vertically polarized radiation, when the wavelength is considerably greater than the diameter of the average tree.

Pulse stretching is not usually a problem with portable beacons because the light weight requirements of the equipment ordinarily preclude the use of a decoder. If there is a decoder, the requirements will be the same as those for permanent ground beacons.

Fig. 20·11.—A satisfactory beacon location. The antenna is higher than near-by vegetation and the sky line is not blocked by trees.

The sky line should be surveyed carefully by the operator before the equipment is set up. Ordinarily no survey with a transit can be made, and a visual inspection must suffice. A good operator will be able to guess with fair accuracy from a visual inspection the directions in which reduced range is to be expected. If certain directions are more important than others, the selection of the site can be made accordingly.

The operator will be able to exercise little control over installation problems. These must be solved in the design of the equipment, which

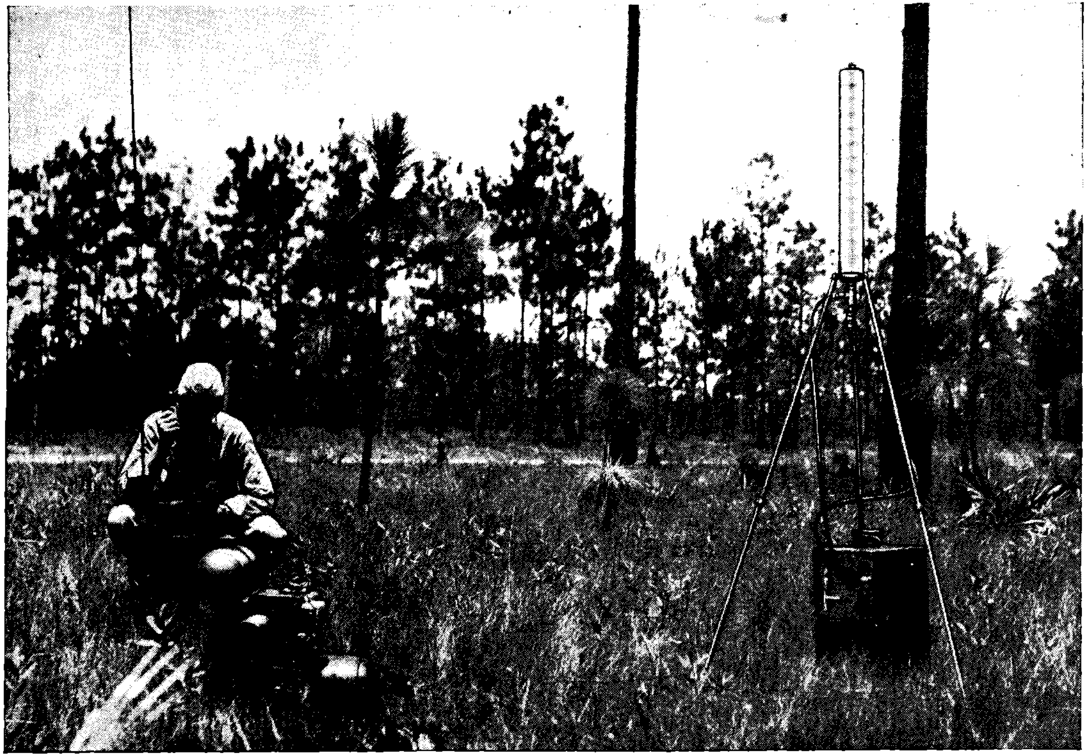


Fig. 20·12.—A moderately good location. The trees are fairly dense, but the sky can still be seen through them. Maximum range will not be greatly reduced.

must provide for assembling, testing, and operating the equipment outdoors under unfavorable weather conditions.

Because little or no auxiliary test equipment will be available, all checking will have to be done with built-in test equipment. Portable microwave beacons have often been so designed that when two beacons were available, one could be used as a signal generator to check the other. A rough field check can readily be obtained in this way.

FIG. 20·13.—A satisfactory beacon location. The sky is visible through scattered trees; range will not be materially reduced.

**20·11. Surface Beacon Systems.**—Previous sections of this chapter have been concerned mainly with beacon systems in which either the interrogator or the beacon was explicitly or tacitly assumed to be airborne. A brief discussion of beacon systems in which both ends of the system are on the surface of the earth is in order because they present some special problems of design, siting, and propagation. Of the three possible surface combinations—ship-to-shore, shore-to-ship, and ground-to-ground—we are particularly concerned with the last. Ship-to-shore and shore-to-ship systems present different, rather special problems.[1]

*Frequency.*—In ground-to-ground systems, relatively short ranges are required. The beacons or interrogators, or both, are usually lightweight and portable. Because of line-of-sight limitations, long ranges—30 miles or more—require considerable elevation at both ends of the system. The signal path is always close to the surface of the earth and often passes through trees or other vegetation. The effect of obstacles of this nature and of resulting diffraction depends on the frequency used.

FIG. 20·14.—A poor beacon location. Underbrush is high and the beacon is in a clump of trees. Range will be materially reduced.

Experiments show that microwave radiation is practically useless in ground-to-ground beacon systems unless there is a clear line-of-sight path for the signal. Signals of lower frequency become progressively better for nonoptical paths, and, at frequencies of 200 Mc/sec or less, a fair amount of transmission through trees and nonoptical transmission

---

[1] A discussion of a shore-beacon–ship-interrogator system can be found in Chap. 19, Secs. 19·10 to 19·13.

becomes feasible. The remaining discussion will be restricted to frequencies below 200 Mc/sec.

*Azimuth Discrimination.*—At the lower frequencies, azimuth determination is most conveniently performed by lobe-switching techniques. Moderate accuracy (±3° to 5°) can be obtained in this way; however, there are often errors. Figure 20·18 shows the distribution of errors of azimuth determinations in 129 fixes in a 175-Mc/sec system with lobe-switched arrays having a half-power beamwidth of about 70°. Less than half of the fixes were over optical paths. The inherent azimuth error under excellent conditions was less than 1°. Some large azimuth errors are due to reflections; these can often be recognized as such by the operator. A more serious type of error is that due to coherent interference

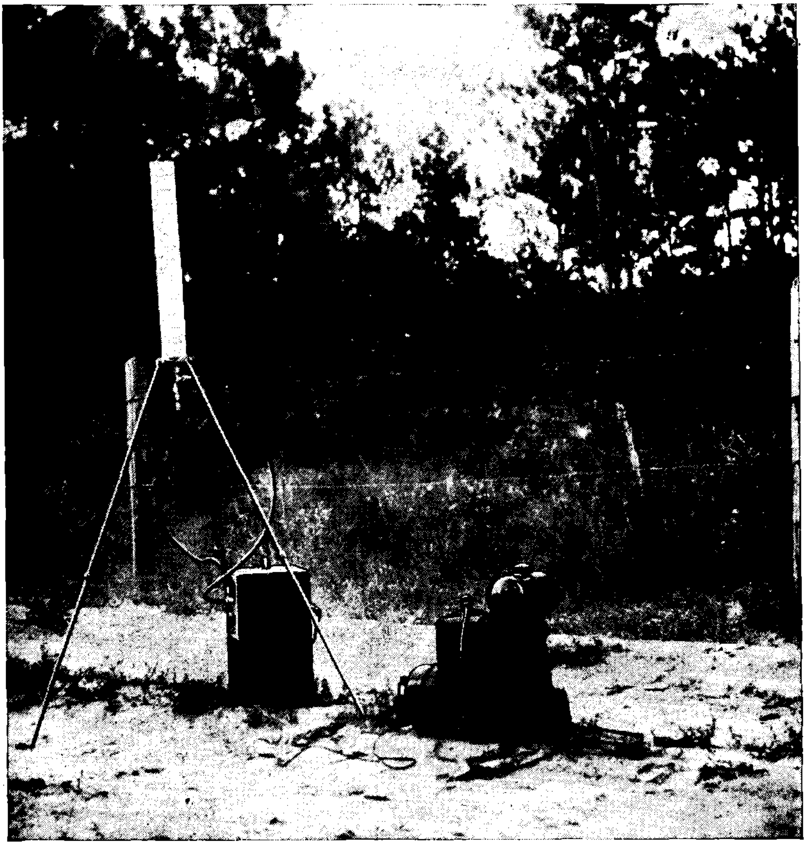


FIG. 20·15.—A poor beacon location. The beacon is in the open, but dense trees near by cut off vision rather completely. Range will be materially reduced.

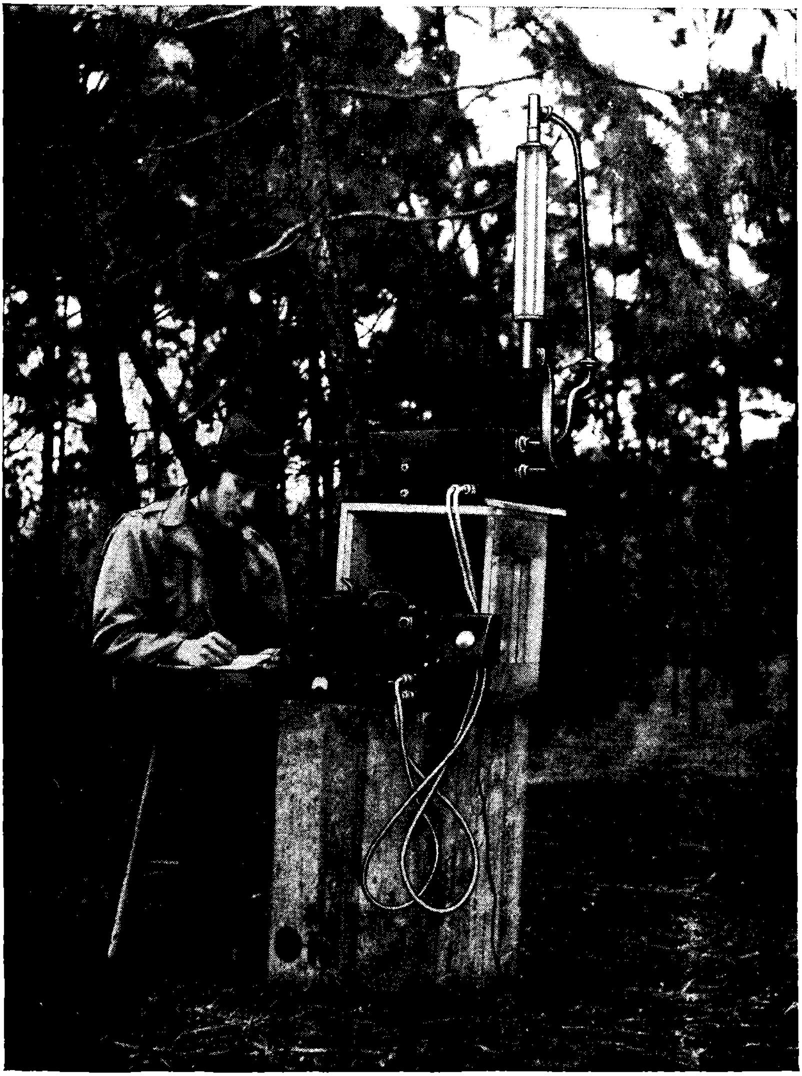


FIG. 20·16.—A poor beacon location. The sky can be seen through the trees in some directions, however. Range will be materially reduced.

**FIG. 20-17.**—An unusable location for beacons. The range will be greatly reduced. Underbrush is higher than the antenna, and the trees are thick.

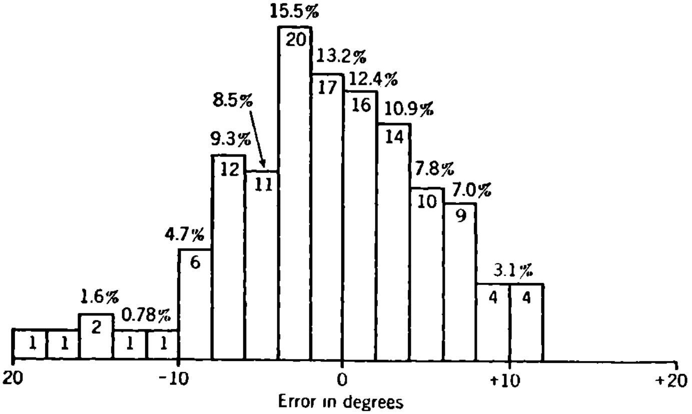


FIG. 20·18.—Distribution of errors in 129 azimuth determinations at 175 Mc/sec by a lobe-switched antenna system having a half-power beamwidth of 70° for each antenna. More than half the fixes were over nonoptical paths.

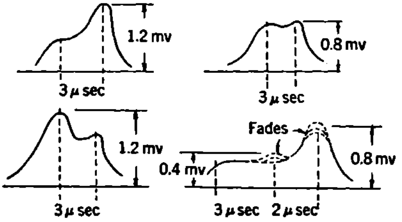
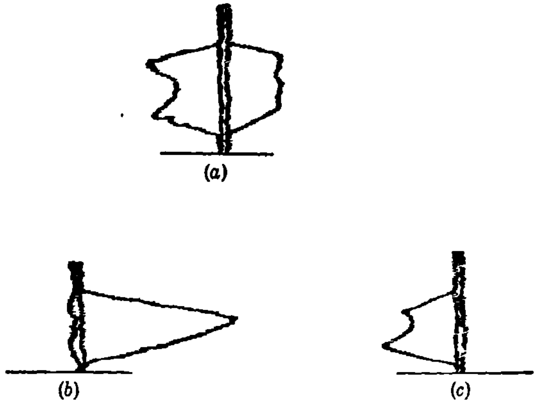


FIG. 20·19.—The change of pulse shape due to coherent interference. A 5-μsec pulse took on the various appearances shown as the path between beacon transmitter and receiver was varied a few feet at a time.


(a)

(b)

(c)

FIG. 20·20.—Effect of coherent interference on pip matching. The shapes of the pulses received by the two antennas are different; matching, and thus azimuth determination, become difficult. (a) Antenna axis pointing at beacon. (b) Antenna axis swung to left. (c) Antenna axis swung to right.

FIG. 20·21.—In passing over this hill, vertically polarized 175-Mc/sec signals were attenuated more than horizontally polarized signals.

between the direct ray and one reflected from neighboring terrain; this is like the interference that normally occurs in transmission over water. However, such interference can change the shape of the pulse, as shown in Fig. 20·19. If the signals displayed on the cathode-ray tube have different pulse shapes, which will be the case if the reflection effects differ for the two switched antennas, matching of amplitudes will then be difficult or impossible (see Fig. 20·20).

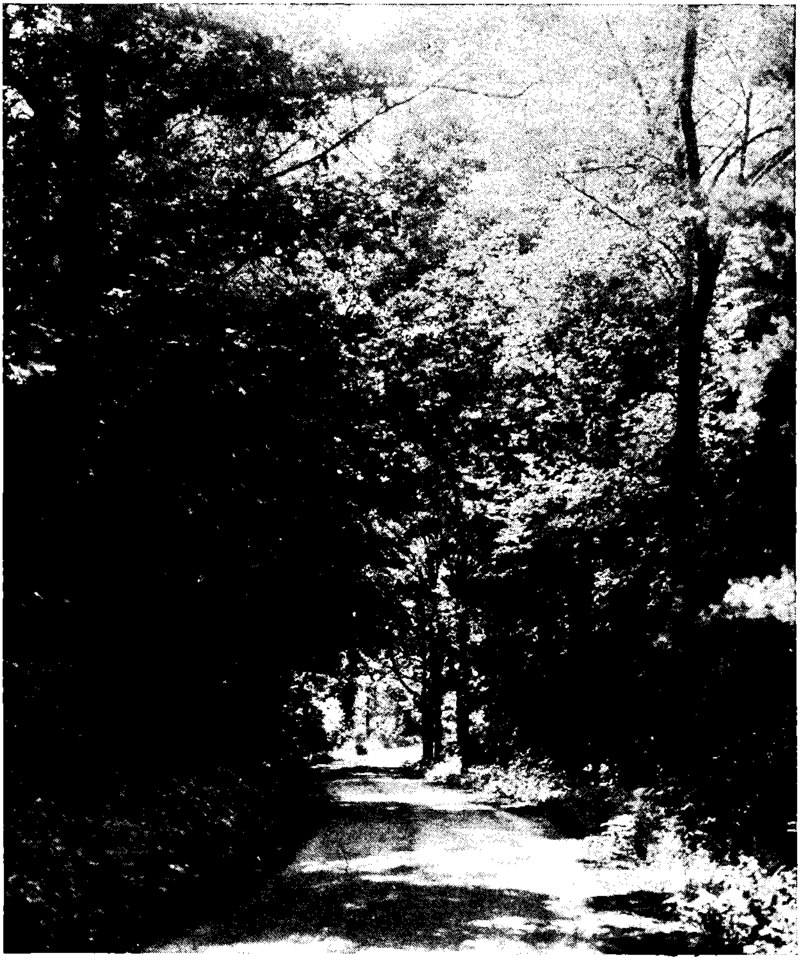


FIG. 20·22.—At 175 Mc/sec, a 100-ft path through these trees attenuated horizontally polarized signals 6 db, and vertically polarized signals 21 db.

Coherent interference can change markedly over very short distances; moving the antenna system a few feet may alter the pulse shapes noticeably. (See Fig. 20·19.)

*Range Errors.*—Range determinations suffer no systematic difficulties in ground-to-ground beacon systems. Reflections can, of course, give false indications; good operating techniques can detect most of these. The most precise ground-to-ground measurements are made with range-only triangulation systems. These are particularly applicable because of the low frequencies necessary and the corresponding difficulty of obtaining good azimuth information. In systems with excess power, azimuth and range determinations can still be made with good accuracy for nonoptical paths, provided the signal strengths are adequate. Except for possible reflections, such determinations are made just as with optical paths. The significance of the measured range is, however, sometimes doubtful when the shortest possible path is longer than a direct optical path.

*Polarization.*—At 175 Mc/sec, a considerable difference exists between the characteristics of vertically and horizontally polarized signals with respect to transmission over nonoptical paths. It is found that vertically polarized radiation is attenuated more by trees than is horizontally polarized radiation. In transmission over obstacles like hills, vertically polarized radiation sometimes gives weaker signals in the diffraction region—perhaps because of the effect of the trees on the hill (see Fig. 20·21).

In the terrain shown in Fig. 20·22, the total attenuation of vertically polarized radiation by about 100 ft of trees was 21 db, whereas horizontally polarized signals passing along the same path were attenuated only 6 db

# APPENDIX A

## AMPLITUDE MODULATION OF PULSES

BY L. A. TURNER

It is well known that an infinite sequence of rectangular pulses of width $\tau$, spaced $T$ seconds apart, can be represented by a Fourier series of harmonic terms having frequencies of 0, $F_0$, $2F_0$, $3F_0$ . . . , where $F_0 = 1/T$. In particular, if zero time be taken at the center of one of the pulses, this series has the form

$$S = \sum_{N=0}^{N=\infty} A_N \cdot \cos 2\pi N F_0 t.$$

The amplitude coefficients $A_N$ are given by

$$A_N = h\tau F_0 \cdot \frac{\sin \pi N F_0 \tau}{\pi N F_0 \tau},$$

in which $h$ is the height of the single pulse.

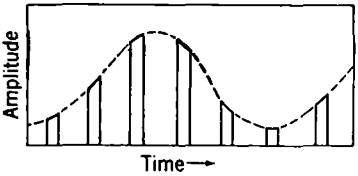


FIG. A·1.—Amplitude-modulated series of pulses.

If the pulses are made to be of an infinitesimal width $\Delta t$, all the amplitudes $A_N$ become equal to $hF_0\Delta t$. It should be noted in passing, since it gives a result needed later, that the pulse of width $\tau$ can be considered as the superposition of a whole set of the infinitesimal pulses appropriately displaced with respect to one another. The corresponding Fourier terms are obtained by combining those of the thin pulses with differences of phase corresponding to the relative displacements of the sets of pulses. In the limit, this becomes the familiar integration encountered in the diffraction of light by a single slit: its result is expressed by the expression given above for $A_N$.

If we now consider the set of pulses of width $\tau$ to be modulated as in Fig. A·1, this can be expressed by multiplying the original rectangular pulse function by a modulating factor of the form $1 + p \cos (2\pi F_m t + \phi)$. Since the rectangular pulse function and the Fourier series are completely equivalent, the series can equally well be multiplied by this factor to get an expression for the modulated waves. Corresponding to each of the orginal Fourier components of frequency $NF_0$, these will now be a triplet of three components at frequencies $NF_0 - F_m$, $NF_0$, and $NF_0 + F_m$ with amplitudes that are respectively $p$, 1, and $p$ times the amplitude of the original component.

If, however, the pulses are still rectangular but modulated in height as in Fig. A·2, the results are slightly different. This case can be handled

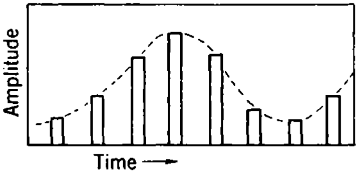


FIG. A·2.—Amplitude-modulated series of pulses.

by starting with a set of infinitesimally thin pulses and applying the modulating factor to them as above to get the same sideband frequencies. The pulses of finite width are now obtained as above by superposition of suitable sets of thin pulses. The amplitude factor becomes

$$hF_0 \frac{\sin \pi F\tau}{\pi F\tau},$$

in which $F$ is the actual frequency of any component whether it be sideband or otherwise. The two sideband components of a given triplet now do not have equal amplitudes, but rather amplitudes that differ in accord with their different frequencies.

# GLOSSARY

**AGC.**—Automatic gain control.

**AI.**—Airborne interception.

**AN/APN-3.**—The SHORAN airborne INTERROGATOR-RESPONSOR.

**AN/APN-19A.**—A modification of AN/APN-19 including double-pulse interrogation coding, with four double-pulse channels.

**AN/APM-53.**—Test equipment for AN/APN-19 beacon.

**AN/APQ-13.**—A 3-cm bombing radar; used by the Twentieth Air Force.

**AN/APS-4.**—A 3-cm airborne radar, used both for sea search and AI.

**AN/APS-15.**—A 3-cm bombing radar; used by the Eighth and Fifteenth Air Forces; MICKEY.

**AN/CPN-2.**—The SHORAN ground beacon.

**AN/CPN-6.**—A high-power ground or shipboard 3-cm beacon.

**AN/CPN-8.**—A medium-power 10-cm ground beacon.

**AN/CPN-17.**—A high-power ground or shipboard 10-cm beacon.

**AN/UPN-1.**—A battery-operated lightweight 10-cm ground beacon.

**AN/UPN-2.**—An a-c-operated lightweight 10-cm ground beacon.

**AN/UPN-3.**—A lightweight a-c-operated 3-cm ground beacon.

**AN/UPN-4.**—A battery-operated lightweight 3-cm ground beacon.

**ASV.**—Air-to-surface-vessel.

**automatic range-tracking.**—The process of obtaining a voltage or shaft rotation proportional to range by using a circuit that locks on and follows a received signal.

**automatic tracking.**—The process of using range data or angular data in such a way as to obtain error signals; these are used to drive devices that keep the system locked to a target.

**beacon.**—See RADAR BEACON.

**blanking gate.**—A circuit used to produce a rectangular pulse when a beacon is triggered; it prevents further triggering for a period of time, called the DEAD TIME.—Also, the rectangular pulse it generates.

**CCB.**—Close control bombing.

**"cat."**—A ground station about which a circular course is flown.

**"cat-mouse" course.**—A method of bombing whereby aircraft fly a circular course around a CAT station and release bombs at a predetermined range from the MOUSE station.

**coding.**—The provision of a special character to interrogation or reply signals.

**cone of silence.**—For a beacon, the approximately conical volume overhead or beneath in which the interrogation or reply link fails because of inadequate antenna coverage. For an airborne scanning INTERROGATOR, the conical volume below not included in the coverage of the antenna scan.

**crystal-video receiver.**—A receiver with a crystal detector and a video amplifier.

**dead time.**—The period following reception of a signal triggering a beacon, during which the beacon cannot again be triggered.

**decoder.**—A device which deciphers a coded signal.

**discriminator.**—A DECODER.

**drift station.**—SHORAN parlance for a CAT beacon.

**fruit.**—In an interrogation display, beacon replies to other INTERROGATORS; they are not synchronized and constitute a type of clutter.

**GCI.**—Ground control of interception.

**Gee.**—A 20-100 Mc/sec hyperbolic navigation system, using pulse transmission.

**Gee-H.**—A British *H*-system for bombing using equipment adapted from the GEE navigation system.

**H-system.**—A beacon bombing system, in which an aircraft equipped with an INTERROGATOR measures precise ranges to two beacons on the ground; the converse of OBOE.

**IFF.**—Identification, friend or foe.

**inquisitor.**—An attachment to a radar set, to enable it to interrogate beacons and obtain their replies.

**interrogation.**—The transmission of pulses designed to TRIGGER a beacon; the pulses triggering a beacon.

**interrogator.**—Any device intended to TRIGGER a beacon by means of r-f pulses transmitted through space.

**interrogation link.**—The communication link whereby an INTERROGATOR triggers a beacon.

**interrogator-responsor.**—A transmitter and receiver especially designed for operation with beacons, rather than as a radar set.

**leaky pipe.**—A slotted coaxial-line or waveguide antenna.

**limiting link.**—That link of a beacon system which determines the reliable range.

**lobe switching.**—A type of azimuth determination.

**Loran.**—A low-frequency hyperbolic navigation system using pulse transmissions.

**Mickey.**—The AN/APS-15 radar.

**Micro-H.**—An American *H*-SYSTEM used for bombing, using the AN/APS-15 airborne bombing radar and AN/CPN-6 ground beacons.

**Minnie.**—The AN/CPN-6 beacon.

**"mouse."**—A reference station from or by which range is measured to an aircraft flying a CAT circle.

**Oboe.**—A beacon bombing system, whereby a beacon-equipped aircraft is guided by two interrogators on the ground; the converse of the *H*-SYSTEM.

**pattern propagation factor** $|F|$.—The ratio of the resultant electric-field amplitude at a point when both reflected radiation and direct radiation reach it to the value due to direct radiation only.

**PDI.**—Pilot's direction indicator, a meter showing a pilot in which direction to fly.

**racon.**—See RADAR BEACON.

**radar beacon.**—A device that, upon reception of a suitable interrogating pulse signal, automatically responds with another pulse signal.

**rate station.**—SHORAN parlance for a MOUSE beacon.

**Rebecca-H.**—A British *H*-SYSTEM used for photo-reconnaissance, adapted from Rebecca-Eureka.

**reply.**—Same as RESPONSE.

**responder beacon.**—See RADAR BEACON.

**response.**—The act of the beacon in emitting a signal when triggered by an interrogation.

**response link.**—The link by which beacon responses are received by an INTERROGATOR.

**ring-around.**—Continuous mutual triggering by two beacons, each of which replies to and triggers the other.

**Rosebud.**—The AN/APN-19 10-cm airborne beacon.

**SCR-584.**—A 10-cm fire-control radar designed for AA use; later adapted for control of aircraft.

**SCR-717.**—A 10-cm airborne ASV radar.

**Shoran.**—A 250-Mc/sec *H*-SYSTEM used for bombing and mapping.

**slow coding.**—A type of response coding in which the response code occupies a time of the order of seconds.

**"squegging" oscillator.**—A self-pulsed oscillator, the pulse duration and repetition rate of which are determined by the oscillator constants.

**squint.**—The dispersion properties of a linear array antenna, which causes the direction of the antenna pattern to vary with frequency.

**STC.**—Sensitivity-time control—see TIME-VARIED GAIN.

**tangential signal.**—A signal which, superposed on noise, appears tangential to the noise on an A-scope.

**time-varied gain.**—An automatic variation of receiver gain with time, usually in such a way as to reduce the gain immediately following a pulse transmission, and bringing it back to its normal value in a predetermined way.

**traffic capacity.**—The ability of an interrogator-beacon system to operate satisfactorily when large numbers of interrogators or beacons are operating.

**transponder.**—See RADAR BEACON.

**trigger.**—To initiate a RESPONSE from a circuit; a signal which initiates such a RESPONSE.

**tripole.**—A set of three dipoles mounted in a circle and excited in phase, to give a pattern uniform in azimuth; used as an element of linear-array antennas.

**TS-143/CPM-1.**—A synchroscope especially designed for microwave beacon testing.

**TVG.**—See TIME-VARIED GAIN.

**wonter.**—A circuit in a double-pulse or multiple-pulse DECODER, which prevents response when an extraneous pulse is present between two pulses with the correct spacing.

**Yehudi.**—A triggering device used in beacon-transmitter AFC systems, which tunes the beacon to the correct frequency when INTERROGATION occurs.

# LIST OF SYMBOLS

In range equations, primed quantities refer to the beacon and unprimed quantities refer to the interrogator.

| | |
|---|---|
| $A$ | Area of antenna; antenna aperture. |
| $A$ | Atmospheric absorption, in decibels. |
| $a$ | Radius of the earth. |
| $B$ | Video bandwidth. |
| $B$ | I-f bandwidth. |
| $b$ | Constant of rectification of a crystal. |
| $C_N$ | Total capacitance of pulse-forming network. |
| $C_s$ | Distributed capacitance of modulator output circuit. |
| $c$ | Velocity of light. |
| $E_B$ | Supply voltage. |
| $F$ | Pattern propagation factor; ratio of electric field to value it would have under free-space conditions. |
| $F$ | Figure of merit of crystal detector. |
| $F_1$ | Noise factor of r-f amplifier. |
| $F_2$ | Noise factor of mixer and i-f amplifier. |
| $F_{12}$ | Over-all noise factor. |
| $G$ | Maximum gain of an antenna, as referred to an isotropic radiator. |
| $h$ | Height. |
| $I_p$ | Pulse current. |
| $i$ | As subscript, refers to interrogation link. |
| $K$ | Loss factor of a transmission line. |
| $k$ | Factor for increasing earth's radius to compensate for atmospheric refraction. |
| $k$ | Boltzmann's constant, $1.38 \cdot 10^{-23}$ joule/degree. |
| $L$ | Transmission line loss in decibels. |
| $L_c$ | Charging inductance for pulse-forming network. |
| $M$ | Reduction from maximum antenna gain to utilized gain, in decibels. |
| $N$ | Number of possible codes. |
| $n$ | Index of refraction of the atmosphere. |
| $n$ | Number of spaces in a range code. |
| $n$ | Noise voltage. |
| $P$ | Number of pips in a range code. |
| $P$ | R-f power. |
| $Q_L$ | Loaded $Q$ of a resonant circuit. |

| | |
|---|---|
| $Q_u$ | Unloaded $Q$ of a resonant circuit. |
| $R$ | Range. |
| $R$ | As subscript, refers to receiving components. |
| $R_c$ | Charging resistor for pulse-forming network. |
| $r$ | As subscript, refers to response link. |
| $S$ | Scanning factor: the ratio of the total angle scanned to the angle over which the beacon is interrogated. |
| $T$ | As subscript, refers to transmitting components. |
| $T_R$ | Interval between pulses; recurrence interval. |
| $W$ | Percent response of a beacon to interrogation. |
| $W_n$ | Percent response of a beacon searchlighted by $n$ interrogators. |
| $Z_N$ | Characteristic impedance of pulse-forming network. |
| $Z_p$ | Impedance presented by primary of pulse transformer. |
| $\lambda$ | Wavelength. |
| $\rho$ | Magnitude of the reflection coefficient of the earth. |
| $\sigma$ | Phase difference between direct and reflected range. |
| $\tau$ | Ratio of beacon dead time to the nominal interrogator repetition period. |
| $\phi$ | Phase change on reflection. |

# *Index*