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RADIOTRONICS

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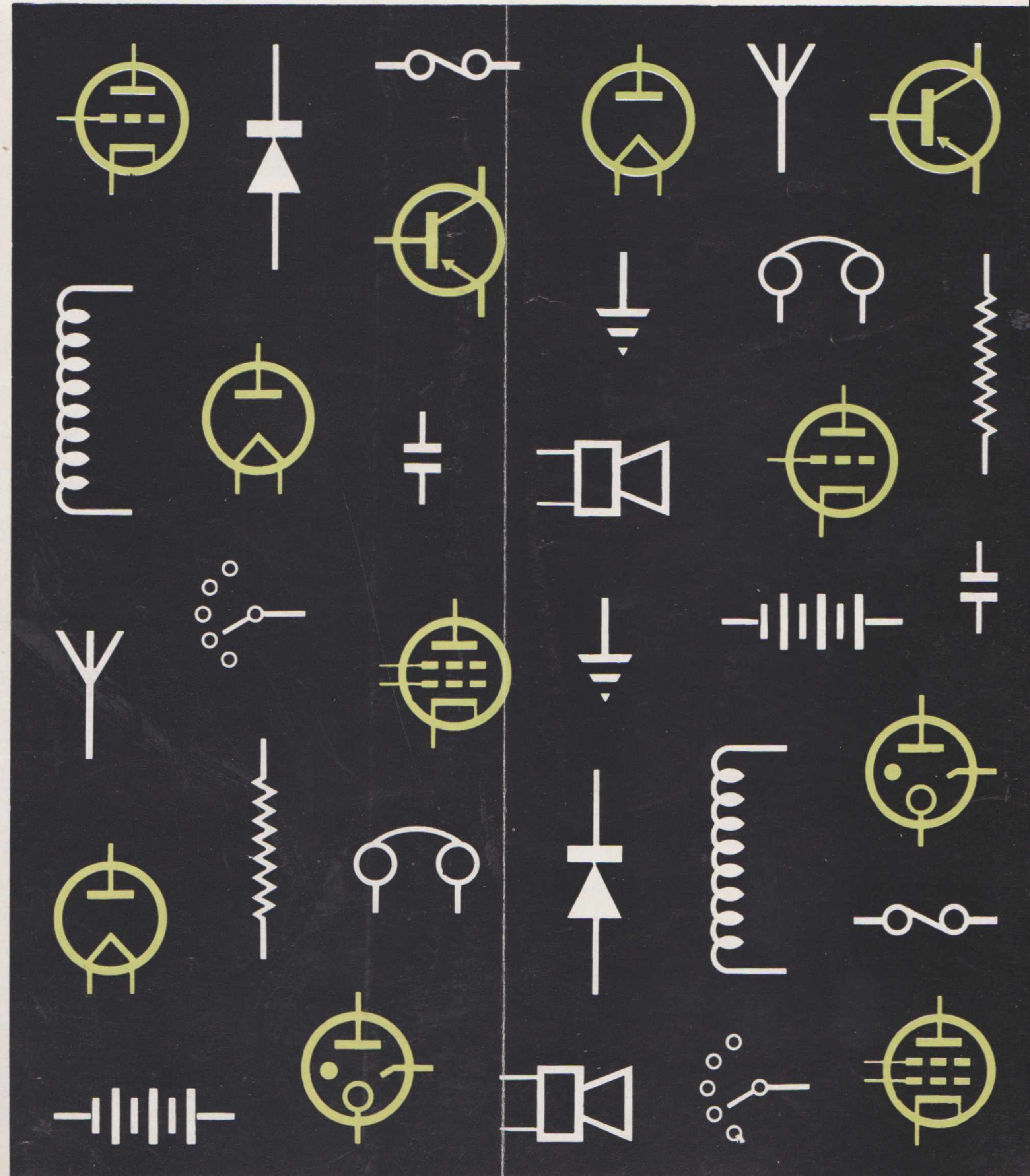

AN AWV PUBLICATION

RADIOTRONICS

Vol. 26, No. 6, 1961

Editor, Bernard J. Simpson

IN THIS ISSUE

The backward-wave oscillator is similar to the forward-wave oscillator, or travelling wave tube as it is more commonly known. Both depend on the interaction between a stream of electrons and an electromagnetic wave travelling on a slow-wave structure. *The difference is that in the backward-wave unit,* the direction of power flow is opposite to that of the beam.

An all-purpose system open to every nation is proposed by RCA.

New releases mentioned this month are: 2N1768, 2N1769, 40-watt n-p-n silicon mesa transistors; 1N3193-1N3196, silicon rectifiers with axaial leads; 2N1700-2N1703, silicon power transistors; 7580, multiplier phototube; 7801, 7870, conduction-cooled cermolox beam power valves; E710, bi-stable direct-view storage tube.

This is an interchangeability guide relating some of the more-common special quality valves to their commercial prototypes. It is intended to assist in the identification of the special quality types.

This is the concluding part of a two-part article dealing with the design of equipment using transistors of this type. The article ends with some typical circuit applications of silicon vhf transistors.

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An Introduction to

"O"-Type Backward-Wave Oscillators

By

J. R. BAGNALL

(English Electric Valve Co. Ltd.)

The "O" type backward-wave oscillator or "O" Carcinotron* is a microwave oscillator which can be tuned electronically and continuously over a very wide frequency range. Several such oscillators having continuous voltage-tuning of an octave or more are now available commercially, and tubes have been described in the literature which can be voltage-tuned over a frequency range of up to 5 : 1.[1] [2] This characteristic of very wide electronic tuning range has made the BWO a very important addition to the existing range of microwave tubes, and a considerable amount of interest and attention has been devoted to it since it was first described in 1952.[3]

Backward-wave oscillators depend for their operation on the interaction between a stream of electrons and an electromagnetic wave travelling on a slow-wave structure. They are very similar therefore to the now conventional travelling-wave tube or forward-wave amplifier, but with one very important difference. In the forward-wave amplifier the radio-frequency energy flow is in the same direction as the electron beam. In the backward-wave tube, however, the power flow is in a direction opposite to that of the beam. Fig. 1 shows schematically the difference between the two types of tube. As will be seen from the diagram, the output of the backward-wave tube is taken from the gun end of the tube at the entry to the slow-wave circuit. Because of the opposite directions of the electrons and the energy flow the interaction between the two is regenerative, and given the proper conditions self-sustaining oscillations result.

The purpose of this article is to explain the principles of operation of this class of oscillator. It is divided into two main sections. The first describes the type of slow-wave circuit used and the nature of the waves propagating on such a circuit. The second describes the nature of the interaction between electrons and the waves and the resulting oscillations. Throughout the article emphasis has been placed on physical principles, and the mathematics have been restricted to the minimum necessary to derive the numerical expressions required for a fundamental understanding of the behaviour of these oscillators.

The Slow-Wave Structure and the Waves

In order to obtain strong interaction between a beam of electrons and an electromagnetic wave the velocity of the electrons must be approximately equal to the phase velocity of the wave. This is explained in detail below in the section dealing with the interaction mechanism. Normally, however, the velocity of electromagnetic waves is approximately equal to the velocity of light, a velocity which is unattainable for electrons accelerated by voltages of the order normally encountered in this type of oscillator. Some slowing down of

*Trade Mark of Cie Générale de Télégraphie Sans Fils, Paris.

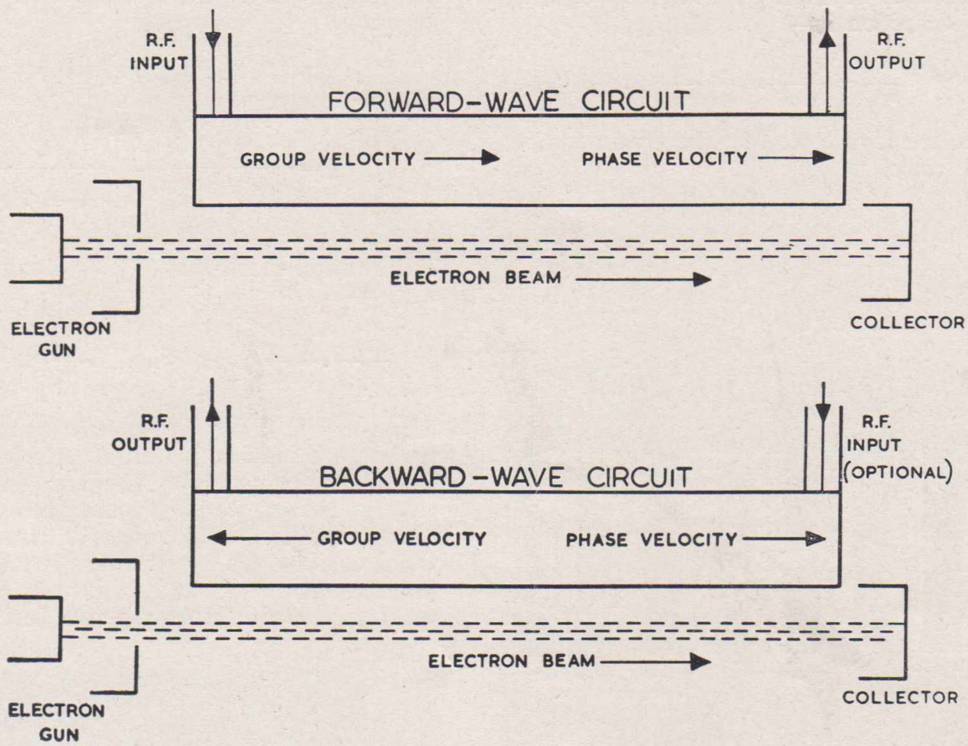


Fig. 1 — Schematic diagrams of a forward-wave amplifier (Travelling-Wave Tube) and a backward-wave amplifier.

the wave is therefore necessary. This slowing-down process can be accomplished in two ways. One method is to make the wave travel in a dielectric medium and the other is to make the wave travel along a slow-wave structure or delay line. It is only the slow-wave structure that is of practical interest.

Two types of delay line are commonly used in the construction of backward-wave tubes. These are the helix and the interdigital line. The present discussion is limited to the interdigital line although much of what is said applies to delay lines in general.

The interdigital delay line is shown diagrammatically in Fig. 2. It consists essentially of two interleaved comb-like conductors as indicated. In the following analysis it will be regarded as a folded parallel-plate transmission line. An electromagnetic wave is assumed to travel in a serpentine path between the fingers as shown by the dashed line in Fig. 2, and its velocity is assumed to be equal to the velocity of light. The electric field configuration along the line is represented in Fig. 3. It is also assumed that fringing fields and discontinuities at the bends of the line are negligible and that the line is lossless. For further simplification the analysis neglects the effect of an electron beam on the electric field.

This type of delay line is a periodic system of periodicity p. That is to say, the wave configuration repeats itself every time it travels a distance p. As a result of this the fields at any point along the line at a distance z differ from those at a distance z + p merely by a constant. If θ_o is the phase change that occurs in the wave in travelling from z to z + p, we can write—

$$F(z + p) = F(z)e^{-j\theta_o} \quad . \quad . \quad . \quad . \quad (1)$$

Simple inspection of this equation reveals that at any given frequency the relation is satisfied not only by θ_o but by many angles θ_m where—

$$\theta_m = \theta_o + 2m\pi \quad . \quad . \quad . \quad . \quad (2)$$

and m can have any integral value, positive or negative. Associated with these phase angles θ_m there are phase constants β_m and phase velocities V_{pm} given by—

$$\beta_m = \frac{\omega}{V_{pm}} = \frac{\theta_o + 2m\pi}{p} \quad . \quad . \quad (3)$$

Or, rearranging for phase velocities—

$$V_{pm} = \frac{\omega p}{\theta_o + 2m\pi} \quad . \quad . \quad (4)$$

Another quantity of interest is the group velocity, which by definition is—

$$\frac{1}{V_g} = \frac{d\beta_m}{d\omega} \equiv \frac{1}{p}\,\frac{d\theta_o}{d\omega} \quad . \; . \; . \; (5)$$

From equation (5) we see that the group velocity is independent of m.

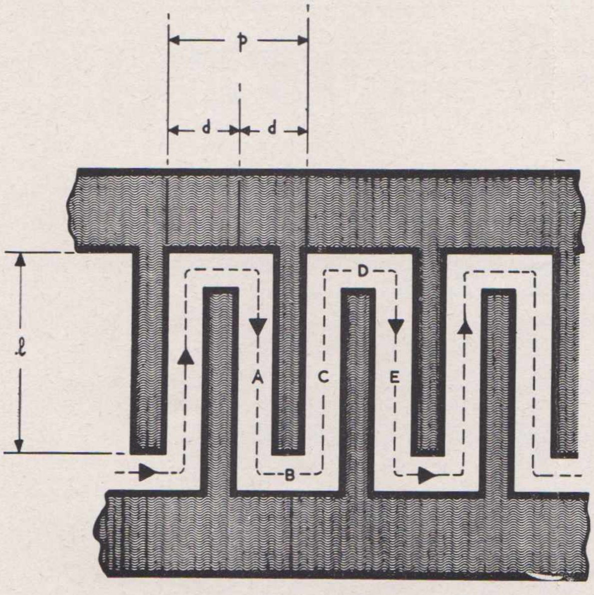


Fig. 2 — The interdigital line.

Equations (4) and (5) show therefore that a periodic structure such as the interdigital line will propagate an infinite number of waves. *At a particular frequency all the waves of this infinite series have different phase velocities which may be either positive or negative but all travel with the same group velocity.* Waves in this series are called *space harmonics* or *Hartree harmonics*. They are analogous to the Fourier components of a periodic curve with sharp corners or discontinuities. That the space harmonics all have the same group velocity should not be surprising because they are all components of a single mode of energy propagation.

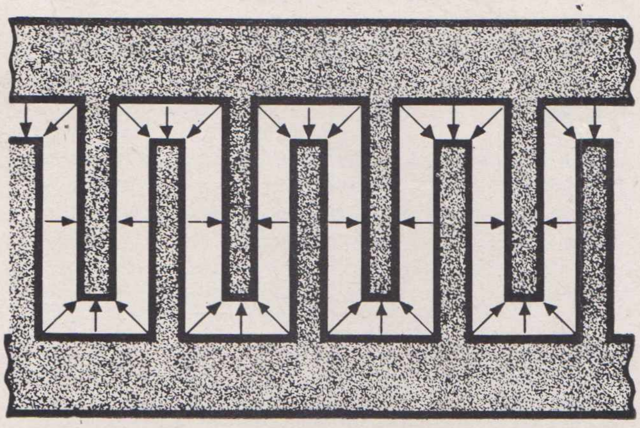


Fig. 3 — Electric field configuration.

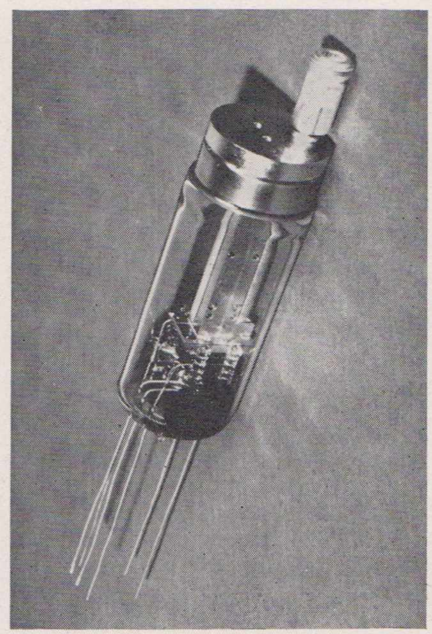
The expression for group velocity may also be written

$$V_g = \frac{1}{(d\beta_m/d\omega)} = \frac{V_{pm}}{1 \, \frac{\omega}{V_{pm}} \cdot \frac{dV_{pm}}{d\omega}} \quad . \; (6)$$

This relation shows that (V_g/V_{pm}) can be negative when—

$$\frac{\omega}{V_{pm}} \cdot \frac{dV_{pm}}{d\omega} > 1 \quad \text{i.e.} \quad \frac{dV_{pm}}{df} > \frac{V_{pm}}{f} \quad . \; (7)$$

Frequency is always a positive quantity in this sense: the above relation therefore requires dV_{pm}/df to be positive. Therefore *when the phase velocity increases with frequency, the group and phase velocities are of opposite sign.* A circuit which exhibits this characteristic is said to possess *negative dispersion*, and it is this characteristic which is vital for backward-wave interaction.



Vacuum envelope of typical EEV Backward-Wave Oscillator, showing internal construction.

To obtain a more detailed knowledge of the interdigital structure we require to know more about the phase velocities of the various space harmonics. This involves simply a knowledge of the phase angle θ_o. Referring to Fig. 2, the phase change in the wave travelling from A to E along the path ABCDE is θ_o and is given by—

$$\theta_o = \frac{2\pi(21 + p)}{\lambda_o} \quad . \; . \; . \; . \; . \; (8)$$

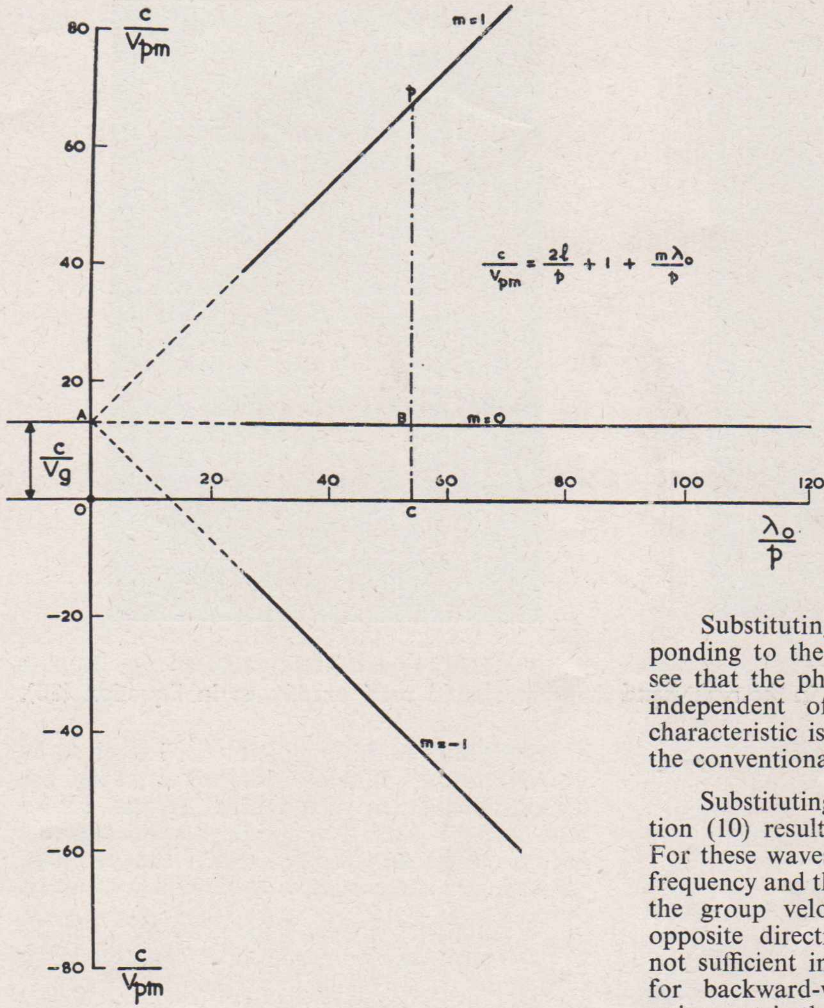


Fig. 4 — Dispersion curves for an interdigital line: l = 6p.

Substituting this value of θ_o in equation (3) we have—

$$\beta m_p = \frac{\omega p}{V_{pm}} = \frac{2\pi(21 + p)}{\lambda_o} + \frac{2m\pi}{p} \quad . \quad (9)$$

or rearranging—

$$\frac{c}{V_{pm}} = \frac{2l}{p} + 1 + \frac{m\lambda_o}{p} \quad . \quad . \quad . \quad (10)$$

Equation (10) shows very clearly the dispersion characteristics of the interdigital line: they are plotted in Fig. 4 for values of m = 0 and ±1. It is interesting to note that the intercept of all the characteristics on the (c/V_{pm}) axis is the same. This intercept can be shown to be equal to (c/V_g). Considering any point P on any characteristic—

$$OA = CP - BP = \frac{c}{V_p} - \frac{\lambda_o}{p} \cdot \frac{d(c/V_p)}{d(\lambda_o/p)}$$

$$= \frac{c}{V_p} - c\lambda_o \frac{d(1/V_p)}{d\lambda_o} \quad . \quad . \quad . \quad (11)$$

This is equivalent to c/V_g : cf. equation (6).

Substituting m = 0 in equation (10), corresponding to the fundamental of the structure, we see that the phase velocity V_{po} is positive and is independent of frequency. This non-dispersive characteristic is the requirement for the circuit in the conventional forward-wave amplifier or TWT.

Substituting negative integers for m in equation (10) results in a series of backward waves. For these waves the phase velocities increase with frequency and therefore, as we have already shown, the group velocity and phase velocities are in opposite directions. This property, however, is not sufficient in itself for the circuit to be useful for backward-wave interaction. A further requirement is that the amplitude of the backward-wave component of interest shall be large enough to present an impedance which will produce adequate interaction between the beam and the circuit.

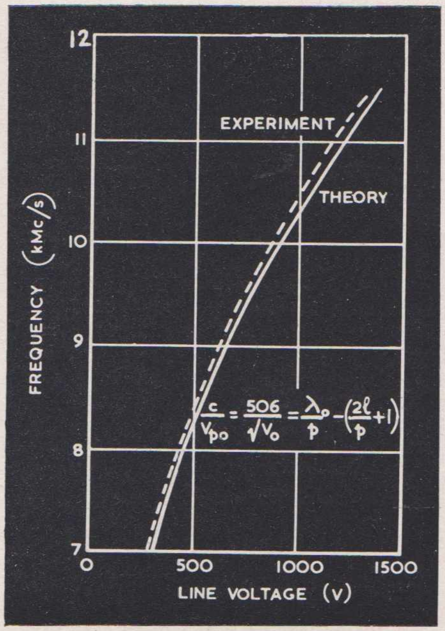
It can be shown that the component with the largest amplitude is the one with the greatest phase velocity. Inspection of equation (10) shows that the backward wave with the largest phase velocity is given by m = −1. The phase velocity of this component is given by—

$$\frac{c}{V_{p(-1)}} = \frac{2l}{p} + 1 - \frac{\lambda_o}{p} \quad . \quad . \quad . \quad (12)$$

At the beginning of this section it was stated that for strong interaction to occur between a beam of electrons and a travelling wave, the electron velocity must be approximately equal to the wave phase velocity. If electrons are accelerated through a potential difference V_o the final velocity of the electrons u_o is given by—

$$\tfrac{1}{2}mu_o^2 = eV_o \quad . \quad . \quad . \quad . \quad (13)$$

where e and m are the charge and mass of the electron.



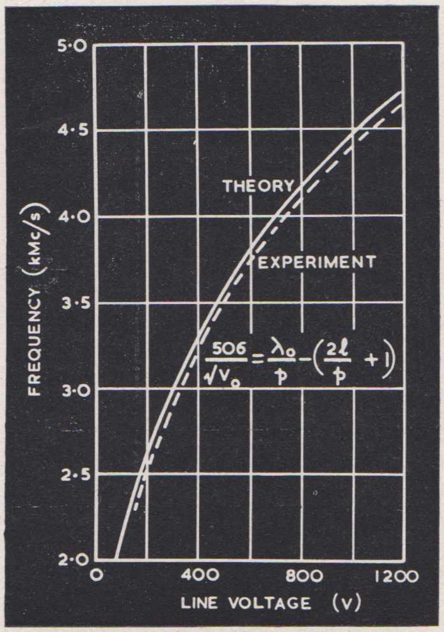


Fig. 5 — Tuning characteristics of backward-wave oscillators; LEFT, type N1034; RIGHT, type N1010. CORRECTION: the equation shown in the right-hand diagram should read exactly as in Equation (15).

The condition for interaction between electrons and waves is therefore—

$$\frac{c}{u_o} \equiv \frac{506}{\sqrt{V_o}} = \left|\frac{c}{V_{pm}}\right| = \left|\frac{2l}{p} + 1 + \frac{m\lambda_o}{p}\right| \quad (14)$$

For the component giving the strongest interaction we have—

$$\frac{506}{\sqrt{V_o}} = \left|\frac{c}{V_{p(-1)}}\right| = \frac{\lambda_o}{p} - \left(\frac{2l}{p} + 1\right) \quad \ldots(15)$$

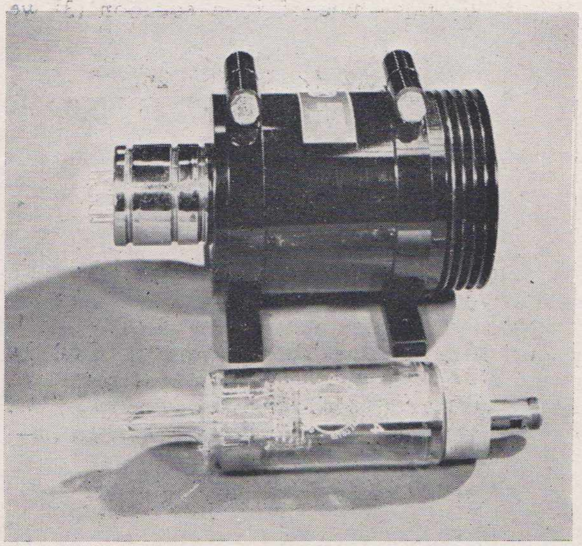
This equation gives the tuning characteristic of a valve employing the first reverse space harmonic. Theoretical curves for valves of the types N1034 and N1010 are plotted in Fig. 5 and actual tuning curves are also shown for comparison. Considering the simplicity of the model and the analysis, the agreement between the actual and predicted curves is remarkably good.

It is at this stage, however, that our simple model fails. It fails completely to predict the total tuning range which can be obtained. The tuning range of such a device is restricted by the allowed variation of θ_o. The restrictions on this angle are $-\pi \leqslant \theta_o \leqslant 0$. Substitution of these values in equation (8) gives for the tuning range $4l + 2p \leqslant \lambda_o \leqslant \infty$. Such a range is unfortunately never realized in practice. The failure of the analysis is inherent in the original assumptions of no fringing fields, no discontinuities and the absence of the effects of the beam.

Also it is not possible to calculate the coupling impedance of the structure using this simple model.

Nevertheless this simple model has given an insight into the nature of backward waves and has demonstrated their plausibility. It has shown that wave components can propagate with opposite group and phase velocities and it has predicted the tuning characteristics with reasonable accuracy.

Having seen some of the elementary details of backward-wave propagation we can proceed to discuss the nature of the interaction between these waves and an electron beam.



EEV Backward-Wave Oscillator type N1034, showing the vacuum envelope and the completed tube mounted in a permanent-magnet focusing unit.

The Interaction Mechanism

In order to further justify equations (14) and (15) we will now explain why the electron velocity and phase velocity of the wave must be approximately equal for interaction to occur. This can readily be seen with the help of the diagram in Fig. 6 which represents an electron beam travelling near a slow-wave circuit. To an observer travelling with a wave on the circuit the electric field has the distribution shown by the dashed lines. The arrows show the direction of the periodic forces acting on the electrons.

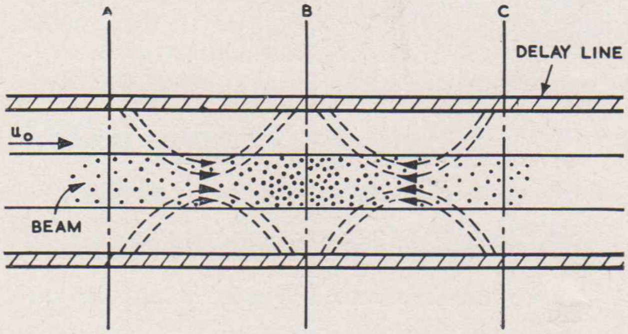


Fig. 6 — Velocity modulation of an electron beam by a wave travelling on a delay line.

If the electron velocity is either much greater than or much less than the wave phase velocity the electrons will experience a large number of wave-troughs and crests. On average the electrons will be accelerated and decelerated for the same length of time. They will therefore on average give up as much energy as they receive, and no net transfer of energy will occur.

If the electron velocity is exactly equal to the phase velocity of the wave the electrons will travel along with the wave. Referring to Fig. 6, electrons in the region AB will be accelerated towards B and therefore will absorb energy at the expense of the wave. Similarly electrons in the region BC will be decelerated towards B and will give up kinetic energy to the wave. A bunch of electrons will therefore form at B, the point of zero field. On average the same number will be accelerated as were decelerated and again there will be no net transfer of energy.

If the electron velocity is slightly in excess of the wave phase velocity the electrons will slowly overtake the wave. Bunching of the electrons will again occur but in this case because of the slight excess velocity of the electrons the bunch will form in advance of B in the region BC. That is to say the bunch forms in a region of a retarding field. We now have more electrons in the regions of a retarding field than there are in an accelerating field. This means that there will be a net transfer of energy from the electrons to the wave. Electrons lose kinetic energy in this process and the amplitude of the wave increases. As this amplitude increases the greater its effect on the electrons and the faster it will increase.

If the electron velocity is slightly less than the wave velocity, bunching will occur in the region AB slightly behind B. As a result there will be a transfer of energy from the wave to the electrons and the amplitude of the wave will decrease.

The most favourable interaction will occur therefore when the electrons travel slightly faster than the wave. And we can now understand more clearly why the backward-wave tube is a voltage-tunable device. The phase velocity of the first reverse harmonic of the interdigital line increases with frequency. If the frequency changes, the electron velocity must be changed to keep in step, as it were, with the wave for maximum interaction. Or in other words the frequency for maximum interaction changes when the electron velocity changes. This device is regenerative and therefore self-sustained oscillations which are voltage-tunable can result. More will be said about the regenerative nature of the device at a later stage.

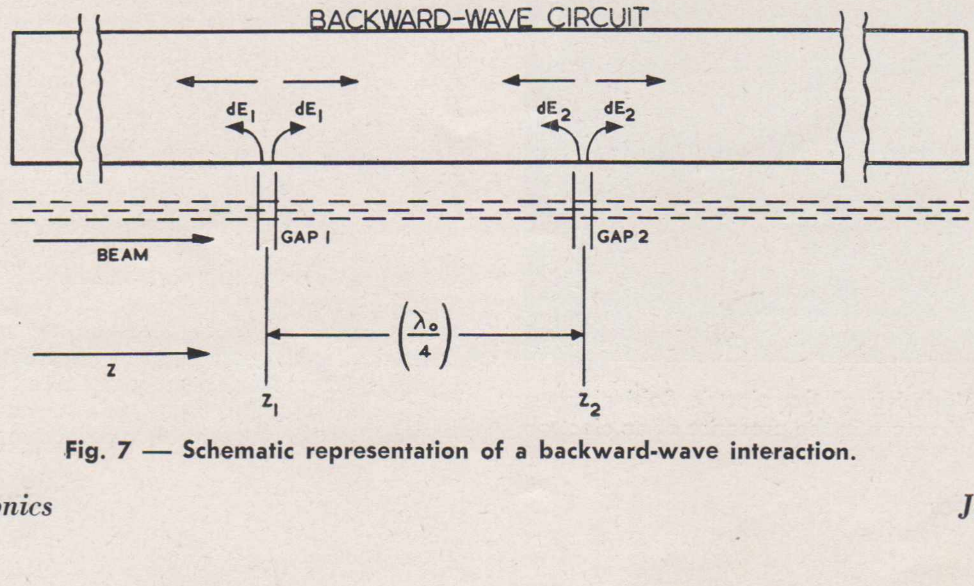
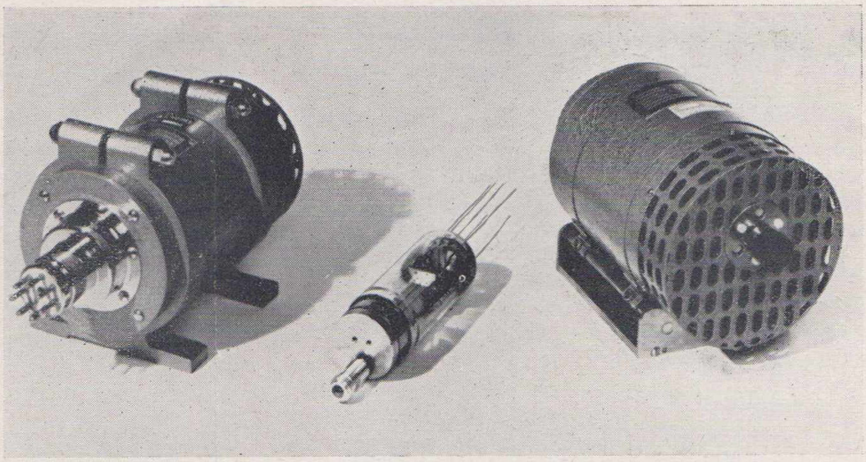


Fig. 7 — Schematic representation of a backward-wave interaction.



On the left is seen a backward-wave oscillator mounted in a permanent-magnet focusing unit. The assembly shown at the right is an electro-magnetic focusing unit. In the centre is a vacuum envelope.

In all the preceding analysis we have always assumed the existence of a wave on the circuit. How is such a wave excited or induced? To explain this we will assume the presence of a radio-frequency current flowing near a backward-wave circuit and examine the reaction. Fig. 7 represents such a system. The system is considered as being divided into a large number of infinitely small gaps of length dz. At each gap the rf current will induce two very small waves of equal magnitude but travelling in opposite directions: these are represented as $\overrightarrow{dE}$ and $\overleftarrow{dE}$. It will be shown that on the backward-wave circuit all the waves $\overleftarrow{dE}$ augment each other while all the waves $\overrightarrow{dE}$ almost cancel.

We will consider these gaps taken in pairs a quarter-wavelength apart (i.e. $Z_2 - Z_1 = \lambda_o/4$). Examining the waves travelling to the right we see that the current will advance in phase by 90° in travelling from Z_1 to Z_2. At Z_2 it will induce two waves $\overrightarrow{dE_2}$ and $\overleftarrow{dE_2}$: consequently $\overrightarrow{dE_2}$ will be advanced in phase by 90° with respect to dE_1 at Z_1. Now $\overrightarrow{dE_1}$ will decrease in phase by 90° in travelling from Z_1 to Z_2: (the direction of power flow is associated with a decrease in phase). At Z_2 we therefore have $\overrightarrow{dE_2}$ and the resultant wave $\overrightarrow{dE_1}$ which has travelled from Z_1 to Z_2. These two components will be 180° out of phase and will therefore tend to cancel. Cancellation is not complete because the current increases in amplitude in travelling from Z_1 to Z_2.

Similarly the induced field waves $\overleftarrow{dE}$ travelling in the opposite direction to the beam will add. The current at Z_2 is 90° ahead in phase of the current at Z_1 and therefore $\overleftarrow{dE_2}$ is 90° ahead in phase of $\overleftarrow{dE_1}$. However, $\overleftarrow{dE_2}$ decreases in phase by 90° in travelling from Z_2 to Z_1 and thus arrives at Z_1 in phase with $\overleftarrow{dE_1}$.

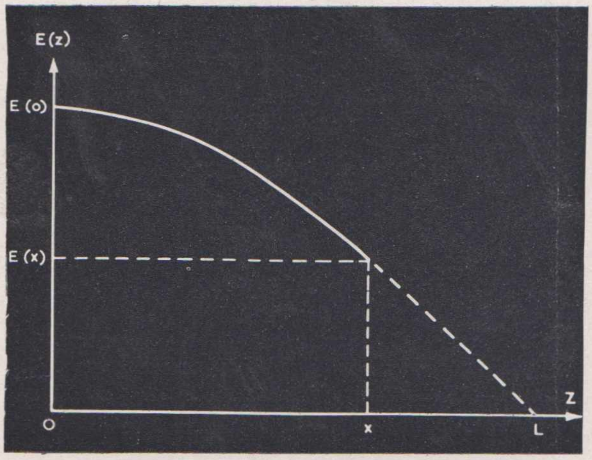


Fig. 8 — Variation of the electric field along a slow-wave structure in the presence of an electron beam.

These small field waves dE add to give a field configuration along the structure which is approximately a cosine distribution. If the field at a point along the structure $z = x$ is represented by E(x) the field distribution along the structure in the presence of an electron beam is as shown in Fig. 8. Consequently if a signal is applied at $z = x$ this signal will become amplified in the direction opposite to the beam. If the field at the entry of the structure is E(o) the gain of the signal appearing at $z = 0$ is given by—

$$\text{Voltage gain} = \frac{E_o}{E(x)} \quad . \quad . \quad . \quad (16)$$

When E(x) is zero the gain in infinite. This occurs at some value of z shown as L. That is to say output can be obtained without any input, or in other words waves will exist in the absence of any applied field, and the system is capable of producing self-oscillations. Physically this means that the coupling between the circuit and the beam becomes sufficient to produce the required modulation in the beam without the aid of an externally applied field. The initial impetus to this reaction is given by the noise fluctuations in the beam.

That this system is regenerative is implicit in what has already been said. In the backward-wave oscillator the induced circuit waves carry energy in the opposite direction to that of the beam: the device therefore has its own built-in feedback mechanism and is regenerative. Also because the current travels at the same velocity as the circuit waves the net phase shift around any closed loop will be zero or some multiple of 2π. For any arbitrary feedback loop within the device, the phase requirement for oscillation is thus automatically satisfied by the condition for maximum backward-wave interaction.

Summary

Backward-wave oscillators are voltage-tuned microwave oscillators. They rely for their operation on the interaction between an electron beam and a travelling wave whose group and phase velocities are in opposite directions. This article has shown that a periodic structure such as the interdigital line can propagate such waves and that interaction occurs when the electron velocity is slightly in excess of the wave phase velocity. Oscillations which are regenerative in nature can occur when the coupling between the circuit and the beam is large enough to produce sufficient modulation of the beam.

The treatment of the Backward-Wave Oscillator given here has been such as to emphasize the physical concepts and mechanisms involved rather than to aim at exactness. A more exact treatment can only be undertaken with the aid of refined mathematical analysis. However the sheer quantity of mathematics involved often tends to obscure the physical principles underlying the overall behaviour. A short bibliography of the more exact treatment is given below.

Acknowledgment

Acknowledgment is made to the Managing Director of the English Electric Valve Company Ltd., for permission to publish this article.

REFERENCES

[1] Sullivan, J. W., " A Wide-Band Voltage-Tunable Oscillator ", Proc. I.R.E., **42**, November, 1954.
[2] Harman, W. A., " Backward-Wave Interaction in Helix-Type Tubes ", Technical Report No. 13 Electronics Research Laboratory, Stanford University, April, 1954.
[3] Tenth I.R.E. A.I.E.E. Electron Tubes Conference, Ottawa, Ontario, June, 1952.

BIBLIOGRAPHY

Kompfner, R. and Walker, N. T., " Backward-Wave Tubes", Proc. I.R.E., **41**, November, 1953.
Pierce, J. R., " Travelling-Wave Tubes ", Van Nostrand, New York, 1950.
Fletcher, R. C., " A Broad-Band Interdigital Circuit for Uses in Travelling-Wave Type Amplifiers ", Proc. I.R.E., **40**, August, 1952.
Johnston, H. R., " Backward-Wave Oscillators ", Proc. I.R.E., **43**, June, 1955.
Warnecke, R. and Guenard, P., " Some Recent Work in France on New Types of Valves for the Highest Radio Frequencies ", P.I.E.E., Part III, **100**, November, 1953.
Heffner, H., " Analysis of the Backward-Wave Travelling Wave Tube ", Proc. I.R.E., **42**, June, 1954.
Palluel, P. and Goldberger, A. K., " The O-Type Carcinotron Tube ", Proc. I.R.E., **44**, March, 1956.
Putz, J. L., Luebke, W. R., Harman, W. A. and Spangenberg, K. R., " Operating Characteristics of a Folded-Line Travelling-Wave Tube ", Technical Report No. 15, Electronics Research Laboratory, Stanford University, June, 1952.
Bernier, J., " Essai de Théorie du Tube Electronique à Propagation d'Onde ", Ann. de Radioélectricité, **2**, January, 1947.
Warnecke, R., " Resultats Obtenus dans le Domaine des Tubes Electroniques ", Ann. de Radioélectricité, **9**, April, 1954.
Leblond, A. and Mounier, G., " Etude des Lignes a Barreaux à Structure Périodique pour Tubes Electroniques U.H.F. ", Ann. de Radioélectricité, **9**, April, 1954.
Hutter, R. G. E., " Beam and Wave Electronics in Microwave Tubes ", Van Nostrand, New York, 1959.

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PUBLICATION RC20

The latest edition of the well-known RCA RECEIVING TUBE MANUAL is now available from the Technical Publications Dept., Amalgamated Wireless Valve Co. Pty. Ltd., Box 2516, G.P.O., Sydney. This publication is a revised and expanded edition, which now contains technical data on more than 760 valves and picture tubes. As in previous editions, this book also contains sections on valve theory, application information and a circuits section. RC20 is 10/- post free.

SATELLITE COMMUNICATIONS

AN ALL-PURPOSE SYSTEM

OPEN TO EVERY NATION IS PROPOSED BY RCA

A new concept in global communications—a single, all-purpose satellite system that would be available to all nations for world-wide telephone, radio, television, telegraph, and data services through two or three relay stations "fixed" in space high above the equator — has been proposed in detail to the Federal Communications Commission (U.S.A.) by the Radio Corporation of America.

RCA told the Commission that such a system could be achieved during the 1960's in a form that would provide communication channels for all international services through their own ground stations, thus eliminating the need for many separate nationally-owned systems. With the RCA proposal was a recommendation to the FCC that space in the radio frequency spectrum be made available for the all-purpose satellite communication system, and that action be taken to encourage its establishment "on a basis that would provide equitable access to all communications carriers".

The proposal, including a detailed description of the all-purpose system and a design for the relay satellites, was contained in a response by RCA and RCA Communications, Inc., to an FCC inquiry relating to the allocation of frequency bands for space communications. In describing the proposed system, based on extensive studies by RCA engineers, the statement pointed out that the concept would meet these major requirements for a practical satellite communications service:

- Service for many nations rather than only one, since the primary purpose is international communication.
- Ability of the satellites to handle signals from many sources at once, in order to permit simultaneous communication between many pairs of ground stations.
- Maximum economy in investment and operating cost, as well as in the use of frequencies.
- Use of a minimum number of satellites in order to avoid cluttering outer space with "useless" orbiting objects.

As described to the FCC, the proposed RCA system would employ synchronous satellite repeaters — space-borne relay stations orbiting 22,300 miles above the equator, where their speed would match the speed of the earth's rotation to keep each satellite effectively "fixed" above one point on the earth's surface.

"A system of this type employing only two satellites could link the major international communications areas of both hemispheres," said the statement. "With three satellites, it would cover every inhabited part of the world, with substantial overlaps."

Each of the satellites would carry relay equipment to receive, amplify, and re-transmit signals between ground stations in various parts of the world, providing a high-capacity link for simultaneous telephone, radio and other desired services. In designs and specifications accompanying the statement, RCA engineers pictured the satellites themselves as relatively small units, each consisting of a slender cylindrical body 13 feet long, with wide fins bearing solar cells to generate power from sunlight, and a dish-shaped antenna at one end, directed constantly toward the centre of the earth.

According to the statement, each satellite could have an initial communications capacity equal to 1,000 two-way telephone circuits, which could be used for telephone, teleprinter, facsimile, or data communications or, alternatively, for a television transmission. It pointed out that all ground stations communicating via the satellite relay would be equipped with directional antennas aimed at the appropriate satellite. The statement also described how the high capacity of the system would permit many services to be carried on at once in accordance with frequency and time allocations established by international agreement and controlled through master ground stations.

Listing the advantages of the proposed system, the RCA statement said: "We believe that this suggested system is feasible today in its electronic requirements, although it will require further development in the techniques of launching and guidance. It uses a minimum frequency band for a given communication capacity, but it is capable of handling a very large communication load. It is highly flexible, providing facilities for either large or small traffic loads among ground stations around the world, and capable of swift rearrangement in accordance with traffic patterns. It requires comparatively simple electronic equipment in the satellite, so that its operation should be relatively reliable over a long period. The equipment would be similar to that in present micro-wave systems, although the components would be specially designed for long life in the space environment."

Emphasizing the flexibility of the system, the RCA statement pointed out that each participating private or national service would be able to establish, own, and operate its own ground station equipment for transmitting and receiving signals of any category through the satellite repeaters. It said:

"This is extremely important in view of the continuing competition among the commercial communications service within the United States, and the desire of the other nations of the world to employ their own national services. With this proposed system, international agreement would be facilitated, since all of the existing services would have individual access to the satellite repeaters from their own ground stations, so that they could continue to conduct their business as they do today."

NEW RELEASES

2N1768, 2N1769

The 2N1768 and 2N1769 are two new 40-watt n-p-n silicon mesa transistors. These new units utilize a hermetically-sealed package having an offset pedestal-and-stud mounting arrangement for positive heat-sink contact. These transistors are designed for such applications as dc-to-dc converters, inverters, choppers, voltage and current regulators, dc and servo amplifiers, and relay-actuating circuits. They feature a very low thermal resistance (junction-to-case) of 4.375° C/watt, a low saturation resistance of 1 ohm and top performance at temperatures up to 200°C.

1N3193, 1N3194, 1N3195, 1N3196

These four new silicon diffused-junction rectifiers are in a tubular, metal case with axial leads, the same small case used in the standard JEDEC TO-1 transistor package. This 0.405-inch-long and 0.240-inch-diameter flangeless case features a single glass-to-metal hermetic seal, and fits easily into standard assembly and mounting procedures. These rectifiers require no derating for operation at ambient temperatures up to 75°C. They have PIV ratings from 200 to 800 volts, with forward currents up to 750 ma.

2N1700-2N1703

The 2N1700 to 2N1703 inclusive are four new silicon power transistors for industrial applications. These new diffused-junction types are especially useful in dc-to-dc converter, inverter, chopper, voltage-and-current regulator, dc amplifier, servo amplifier, power oscillator, and relay-actuator circuits. Capable of operating at case and/or mounting-flange temperatures from —65° to +200°C, these new silicon power transistors feature top performance at high temperatures, dissipation capability up to 75 watts, and the high maximum collector-to-base voltage rating of 60 volts.

7580

The 7580 is a head-on type multiplier phototube having extremely fast rise time in conjunction with high current amplification. It is designed for fast-coincidence work and other very-low-light level applications where wide bandwidth, very small spread in electron transit time, and good resolution are desired. The 7850 is particularly suited for detecting events producing only a few photons. When operated at a supply voltage of 2300 volts, the 7850 has median luminous sensitivity of 6000 amperes per lumen; current amplification of 86,000,000; and anode pulse rise time of 2 x 10^{-9} second. Median pulse height resolution for the 7850 is 8.5 per cent.

7801, 7870

The 7801 and 7870 are two new, very small, conduction-cooled uhf cermolox valves designed for a wide variety of applications in which small size and compactness of equipment is of utmost importance. These beam power valves feature outstanding performance as rf power amplifiers and oscillators, but may also be used as voltage regulators where again space is at a premium. The 7801 and 7870 differ only in heater voltage and current; the 7801 operates at 12.6 volts/0.5 ampere, and the 7870 operates at 6.3 volts/1.0 ampere. At 3000 Mc these new valves can provide 3.2 watts of CW power output with one watt of driver power output, or they can provide a gain of 9 db with 50 milliwatts of driver power output.

E710

A new approach to the design of the storage target assembly has been used in the E710 bi-stable direct view storage tube by EEV. The use of the new technique involves far less risk of damage to the assembly should a high beam current be applied accidentally, and so the associated circuitry can be simplified. This tube will store information indefinitely and allow the continuous or intermittent display of the stored waveform. Two or more signals can be stored and simultaneously displayed and instantly erased when no longer needed. The E710 has a writing gun, electrostatically deflected and focused, which deposits a charge pattern on the storage mesh, the spot diameter being approximately .010". A second gun, which floods the storage mesh continuously, maintains this signal and displays it on the phosphor. Two years ago the first British half tone storage tube, the E702, was announced by EEV and now the E710 bi-stable storage tube makes this a double first in British storage tube technique.

KU-BAND TWT CHAIN

In a paper written with Ralph E. Bridge, Wayne J. Caton described the design of a lightweight travelling-wave-tube amplifier chain operating over the 12,000-15,000-megacycle frequency band. The two RCA engineers were addressing the American Institute of Electrical Engineers in New York. The amplifier chain consists of two ruggedized travelling-wave tubes operated in tandem. Both tubes are periodically focused and have been designed specifically to meet the environmental demands of airborne equipment, said Mr. Caton.

The driver tube in the chain is a developmental 10-milliwatts, 30-db-gain amplifier. The output tube, also in the development stages, produces a minimum power output of 1 watt with a saturated gain of 30 db. According to Mr. Caton, the completely packaged 1-watt tube is approximately 15 inches long and weighs 6½ pounds. The 10-mw tube weighs 4 pounds and is 15 inches long. RF input and output connections on both tubes are made through standard UG-419/U waveguide flanges. External electrical connections are made through silicone-insulated flying leads to permit high-altitude operation.

Explaining the structure of the amplifier chain, Mr. Caton stated that all package parts are potted into place with an epoxy resin to minimize the effects of shock and vibration. Each tube is supported in the package at both ends and potted in the centre with a silicone rubber.

To keep the axial magnetic field of the tubes constant with temperature, the periodic focusing structure is temperature-compensated with nickel-iron alloys. These alloys also reduce the stray magnetic leakage fields usually associated with the focussing structure. This design feature allows the tubes when used in tandem operation, to be arranged very close together.

In tandem operation, the RCA engineer said, these tubes become an extremely high-gain, high-frequency amplifier chain. This chain will provide a minimum power output of 1 watt and a minimum gain of 60 db across the 12,000-15,000 megacycle frequency band. This amount of gain assures a minimum power output of 1 watt for an input drive power of approximately 1 microwatt. The complete chain weighs 10½ pounds and operates with 75 watts of dc power.

PREMIUM AND SPECIAL VALVES

AN INTERCHANGEABILITY GUIDE

Over recent years many new valve types have been introduced for special applications. These special applications include those where the valve is subjected to an environment which places unusual electrical or mechanical stresses on the valve, those where the mode of operation requires special characteristics, and those where a tighter specification is required.

These applications could be respectively exemplified by operation under severe shock and vibration or at high altitudes, use in computing equipment or other on-off applications, and cases where exceptional uniformity of characteristics, low hum level or some other characteristics figure in the specification. There are of course many other circumstances in which it is desirable to use a special-quality valve rather than a normal commercial type.

In many cases the special-quality valve is based on a commercial prototype and is either identical with or very similar to it. This listing, which is not claimed to be complete, has been prepared to assist readers in identifying some of the special-quality types in relation to the commercial prototypes, and in most cases indicates the main points of difference.

Please note that where no difference is noted in the list, this does not necessarily mean that the type and prototype are electrically and mechanically identical. Published characteristics should be studied where necessary to establish the difference. Further data on special-quality valves appeared in "Radiotronics", Vol. 24 No. 8, August 1959, and in the AWV publication "Special Quality Valves", Publication No. SQ-1, 1st edition, June 1960. (3/- post free).

TYPE	CLASS	FEATURES	PROTOTYPE
1612	Heptode	Low microphony	6L7
1620	Sharp CO Pentode	Low microphony	6J7
1621	Power Pentode		6F6
1622	Beam Pentode		6L6
1629	Electron Ray Tube	12.6-volt heater	6E5
1631	Beam Pentode	12.6-volt heater	6L6
1632	Beam Pentode	12.6-volt heater	25L6
1634	Twin Triode	Matched sections	12SC7
5591	Pentode		6AK5
5654	Sharp CO Pentode		6AK5
5656	Sharp CO Pentode	Internal capacitor g2-k	6AK5
5670	Twin Triode	Medium-mu, 350-ma heater	2C51
5691	Twin Triode	Balanced sections, 600-ma heater	6SL7
5692	Twin Triode	Reduced plate dissipation	6SN7
5693	Sharp CO Pentode	Reduced plate, screen dissipations	6SJ7
5725	Gate Pentode	175-ma heater, higher ratings	6AS6
5726	Twin Diode	Balanced sections	6AL5
5727	Gas Tetrode		2D21
5749	Rem. CO Pentode	Controlled characteristics	6BA6
5750	Heptode	Controlled characteristics	6BE6

TYPE	CLASS	FEATURES	PROTOTYPE
5751	Twin Triode	Balanced sections, 175/350-ma heater, lower mu	12AX7
5814A	Twin Triode	Balanced sections, medium-mu 175/350-ma heater	12AU7
5824	Beam Pentode		25B6G
5838	Rectifier	12-volt, 600-ma heater, 10 per cent. higher ratings	6X5GT
5839	Rectifier	26.5-volt, 285-ma heater, 10 per cent. higher ratings	6X5GT
5842	Triode	High gm	417A
5844	Twin Triode	For computer use	6J6
5847	Sharp CO Pentode	For carrier equipment	404A
5852	Rectifier	1200-ma heater, 10 per cent. higher ratings	6X5
5871	Beam Pentode	Derated 10 per cent.	6V6GT
5881	Beam Pentode	10-20 per cent. higher ratings	6L6
5915	Heptode	Derated slightly	6BE6
5930	Power Triode		2A3
5931	Rectifier		5U4G
5932	Beam Pentode	10-20 per cent. higher ratings	6L6
5933	Beam Pentode		807
5965	Twin Triode	Balanced sections	12AV7
5992	Beam Pentode	Derated 20 per cent., 600-ma heater	6V6GT
5993	Rectifier	Different basing, 800-ma heater	6X4
5998	Twin Pow. Triode	Medium mu	421A
6004	Rectifier	Plate caps	5Y3GT
6005	Beam Pentode		6AQ5
6028	Sharp CO Pentode	For carrier equipment	408A
6042	Twin Triode	25-volt, 150-ma heater	6SN7
6045	Twin Triode	350-ma heater	6J6
6046	Beam Pentode		25L6
6057	Twin Triode		12AX7
6058	Rectifier		6AL5
6059	Pentode		6BR7
6060	Twin Triode		12AT7
6061	Beam Pentode		6BW6
6063	Rectifier		6X4
6064	Pentode		6AM6
6066	DD Triode		6AT6
6067	Twin Triode		12AU7
6072	Twin Triode	Low-noise, 175/350-ma heater, higher mu	12AY7
6073	Gas Diode	Shock resistant. $V_S = 185$ volts	0A2
6074	Gas Diode	Shock resistant. $V_S = 133$ volts	0B2
6080	Twin Pow. Triode	Low-mu, shock resistant	6AS7G
6082	Twin Pow. Triode	26.5-volt, 600-ma heater	6AS7G
6087	Rectifier		5Y3GT
6094	Beam Pentode	Different basing, 600-ma heater	6AQ5
6095	Beam Pentode		6AQ5
6096	Sharp CO Pentode		6AK5
6097	Rectifier		6AL5
6098	Beam Pentode		6AR6
6099	Twin Triode	Balanced sections	6J6
6100	Triode	Low microphony	6C4
6101	Twin Triode	Balanced sections, red. plate dissipation	6J6
6106	Rectifier		5Y3GT
6113	Twin Triode	600-ma heater	6SL7GT
6132	Pentode		6CH6
6134	Sharp CO Pentode		6AC7
6135	Triode	Controlled characteristics, 175-ma heater	6C4
6136	Sharp CO Pentode	Controlled characteristics	6AU6
6137	Rem. CO Pentode	Controlled characteristics	6SK7
6180	Twin Triode		6SN7
6186	Sharp CO Pentode	Operation to 400 Mc.	6AG5
6187	Gate Pentode	175-ma heater, higher ratings	6AS6
6188	Twin Triode	High-mu	6SU7WGT
6189	Twin Triode		12AU7

TYPE	CLASS	FEATURES	PROTOTYPE
6197	Power Pentode		6CL6
6201	Twin Triode	Operation to 300 Mc.	12AT7
6202	Rectifier	Reduced ratings	6X4
6203	Rectifier	Different basing, 900-ma heater	6X4
6205	Sharp CO Pentode	G3 on separate pin	5840
6206	Rem. CO Pentode	G3 on separate pin	5899
6265	Sharp CO Pentode	175-ma heater, plate dissipation 2W	6BH6
6336	Twin Pow. Pentode	4.75-amp heater, higher ratings	6AS7G
6384	Beam Pentode	Higher ratings	6AR6
6385	Twin Triode	500-ma heater	2C51
6386	Twin Triode	Remote cutoff, 350-ma heater	2C51
6388	Gas Triode		443A
6394	Twin Triode	26.5-volt, 1.2-amp heater, higher ratings	6AS7G
6417	Beam Pentode	12.6-volt heater	5763
6485	Sharp CO Pentode		6AH6
6486	Gate Pentode	Higher ratings, 250-ma heater	6AS6
6516	Pentode		6AM5
6520	Twin Triode	Balanced halves, 600-volt g-p insul.	6AS7
6528	Twin Pow. Triode	High gm and Ip, 5-amp heater	6AS7
6533	Triode	Low microphony	6247
6626	Gas Diode	$V_S = 165$ volts	0A2
6627	Gas Diode	$V_S = 130$ volts	0B2
6660	Rem. CO Pentode	For mobile use	6BA6
6661	Sharp CO Pentode	For mobile use. Improved heater	6BH6
6662	Rem. CO Pentode	For mobile use	6BJ6
6663	Rectifier	For mobile use	6AL5
6669	Beam Pentode	For mobile use. Improved heater	6AQ5
6677	Power Pentode	For mobile use. Improved heater	6CL6
6678	Triode-Pentode	For mobile use. Improved heater	6U8
6679	Twin Triode	For mobile use	12AT7
6680	Twin Triode	For mobile use	12AU7
6681	Twin Triode	For mobile use	12AX7
6829	Twin Triode	Derated. Longer bulb	12AV7
6830	Gas Diode	$V_S = 185$ volts	0A2
6831	Gas Diode	Flying leads	0B2
6913	Twin Diode	Medium mu	12BH7
6927	Twin Triode	330-ma heater	6J6
6928	Beam Pentode	360-ma heater, reduced ratings	6AQ5
6968	Sharp CO Pentode	Controlled cutoff	6AK5
6973	Beam Pentode	Higher ratings	6CZ5
7025	Twin Triode	Controlled noise, hum	12AX7
7036	Heptode	Derated slightly. Longer bulb	6BE6
7054	Power Pentode	13.5-volt heater, slightly derated	12BY7
7055	Rectifier	13.2-volt heater	6AL5
7056	Sharp CO Pentode	12-15-volt, 150-ma heater	6CB6
7057	Twin Triode	12-15-volt, 175-ma heater	6BZ7
7058	Twin Triode	13.5-volt, 155-ma heater	12AX7
7059	Triode Pentode	12-15-volt, 195-ma heater	6U8
7060	Triode Pentode	Heater-cathode voltage, 120 volts	6AU8
7061	Beam Pentode	13.5-volt, 210-ma heater	12AB5
7079	Twin Triode	For uhf service	6111
7083	Sharp CO Pentode	Higher operating frequency	5702
7105	Twin Pow. Triode	12.6-volt, 1.25-amp heater	6AS7G
7137	Triode	Medium mu	6J4
7167	Sharp CO Tetrode	12-15-volt, 90-ma heater	6CY5
7212	Beam Pentode		6146
7244A	Twin Triode	Medium mu, shorter bulb	6J6
7245A	Triode	Shorter bulb	6J4
7258	Triode-Pentode	13.5-volt, 210-ma heater	6AN8
7320	Beam Pentode		6BQ5

SILICON VHF TRANSISTORS

AN APPLICATION GUIDE

(CONTINUED)

High-level Video Amplifiers

The use of transistors in video amplifiers is complicated by both the low input impedance of transistors and the frequency dependence of this impedance. In general, the response of a typical high-frequency transistor amplifier is more dependent upon input parameters than on output. If any peaking or high-frequency compensation is employed, therefore, it should be concerned primarily with the base-to-emitter circuit rather than the collector circuit.

A simple and effective means of improving the response of the input network is emitter peaking. This type of peaking improves both the linearity and the stability of the amplifier. It also minimizes the effects of varying or different transistor internal parameters on the characteristics of the circuit.

The following analysis is based on the familiar hybrid-pi equivalent network shown in Fig. 10, and also applies to the amplifier shown in Fig. 11. The analysis is divided into three parts : (1) calculation of an equivalent input network, (2) calculation of an output network, and (3) combination of (1) and (2) to provide useful design equations.

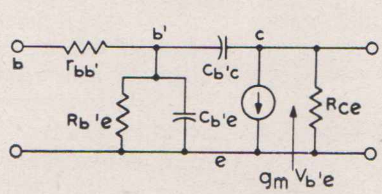


Fig. 10—Simplified hybrid-pi equivalent circuit.

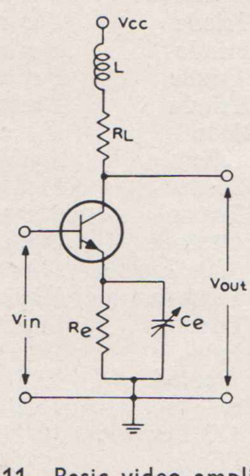


Fig. 11—Basic video amplifier.

If it is assumed that the equivalent base resistance $r_{bb'}$ is part of the generator impedance, the input circuit then appears as shown in Fig. 12. At low frequencies, this input impedance appears as a parallel resistance R_1 and capacitance C_{in} having the following values :

$$C_{in} = \frac{C_1}{1 + g_m R_e} \quad (12)$$

$$R_1 = R_{b'e}(1 + g_m R_e) \quad (13)$$

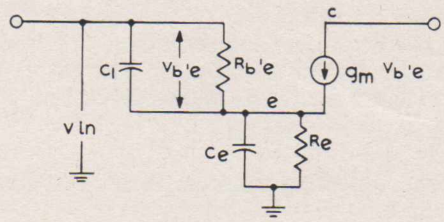


Fig. 12—Equivalent input circuit.

However, C_i must also include the feedback capacitance, which is equivalent to Miller capacitance in electron valves. Therefore, equation (12) may be rewritten as follows :

$$C_i = C_{b'e} + C_{b'c}\, g_m R_L \tag{14}$$

For determination of the time constant and loss of the input network, the effects of $r_{bb'}$ and the generator resistance must be included, as shown in Fig. 13. The equivalent input resistance R_{in} is the parallel combination of $R_{b'e}$ and R_g plus $r_{bb'}$:

$$R_{in} = \frac{R_i\,(r_{bb'} + R_g)}{R_i + r_{bb'} + R_g} \tag{15}$$

where R_g is the signal-source internal resistance.

The time constant of the input T_{in} is the product of R_{in} and C_{in} :

$$T_{in} = R_{in}\, C_{in} \tag{16}$$

There is a small-signal loss in the input network as a result of $r_{bb'}$. This loss $A_{bb'}$ may be computed as follows :

$$A_{bb'} = \frac{R_i}{R_i + r_{bb'}} \tag{17}$$

The analysis of the collector circuit at low frequency involves the parameters shown in Fig. 14. The output voltage V_o is given by

$$V_o = g_m V_{b'e} R_L \tag{18}$$

The input voltage V_{in} is given by

$$V_{in} = V_{b'e} + g_m V_{b'e} R_e \tag{19}$$

The internal gain from base to collector $A_{b'c}$ may be computed as follows :

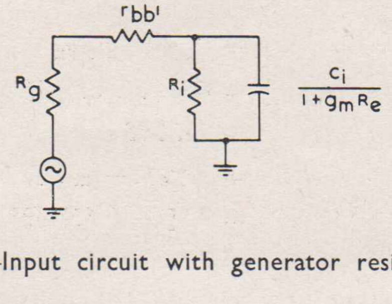


Fig. 13—Input circuit with generator resistance.

$$A_{b'c} = \frac{V_o}{V_{in}} = \frac{g_m V_{b'e} R_L}{V_{b'e} + g_m V_{b'e} R_e}$$

$$= \frac{g_m R_L}{1 + g_m R_e} \tag{20}$$

The total gain A is then given by

$$A = (A_{bb'})\,(A_{b'c})$$

$$= \left(\frac{R_i}{R_i + r_{bb'}}\right)\left(\frac{g_m R_L}{1 + g_m R_e}\right)$$

$$= \frac{g_m R_L R_{b'e}}{R_{b'e}\,(1 + g_m R_e) + r_{bb'}} \tag{21}$$

The value of g_m at 25 degrees Centigrade is approximately given by

$$g_m = 0{\cdot}038\ I_c\ (\text{ma}) \tag{22}$$

The time constant of the collector circuit T_{out} is given by

$$T_{out} = \frac{R_L C_o}{1 + g_m Z_e} \tag{23}$$

where $C_o = C_{ob} + C_{wiring} + C_{load}$

Z_e = parallel combination of R_e and C_e

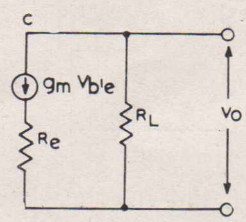


Fig. 14—Equivalent output circuit.

This time constant may be improved by shunt or series peaking. More complex peaking generally will not produce an appreciable improvement in response of the entire amplifier, however, because the input network usually imposes the greatest restriction on bandwidth.

For example, shunt peaking may be employed with an allowable overshoot of 2·5 per cent. In such a case, bandwidth improvement in the collector of approximately 1·7 may be realized, i.e., the output time constant will be reduced by a factor of 1·7, as follows :

$$T_{out} = \frac{R_L C_o}{1{\cdot}7\ (1 + g_m Z_e)} \tag{24}$$

For such shunt peaking of an amplifier, a small inductor L is placed in series with the collector load resistance R_L. This inductor has the following value :

$$L = m\,R_L^2\,C_o \qquad (25)$$

where m is 0·4 for a bandwidth improvement of 1·7 and an overshoot of 2·5 per cent.

If r_{bb}' is neglected, the gain of the amplifier A′ is given by

$$A' = \frac{g_m\,R_L}{1+g_m\,R_e} \qquad (26)$$

It can be seen that the gain of the amplifier is reduced by the factor $(1+g_m\,R_e)$ as compared to that of an equivalent amplifier having either no emitter resistance or a well-bypassed emitter resistance. Because the time constant or bandwidth is improved by exactly the same ratio, the gain-bandwidth product remains constant.

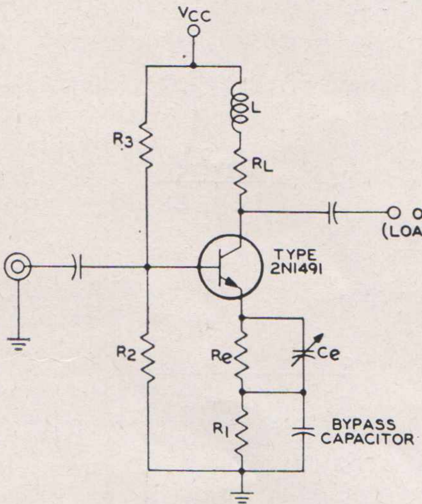


Fig. 15—Video amplifier.

The emitter resistance performs to some extent as a divider to the signal being amplified. The flattest response is achieved when the emitter-circuit time constant T_e equals the time constant of its associated parameters, as follows :

$$R_e\,C_e = [(T_{in})^2+(T_{out})^2]^{\frac{1}{2}} = T_e \qquad (27)$$

The upper cutoff frequency of the amplifier f_{3db} is then given by the following equation for the maximally flat condition :

$$f_{3db} = \frac{1}{2\pi\,T_e} \qquad (28)$$

If some overshoot or undershoot is desired or can be tolerated, the following approximate expression for the upper cutoff frequency may be used :

$$f_{3db} = \frac{1}{2\pi\,T_e \dfrac{C_e}{C_e+\triangle C}} \qquad (29)$$

Over a small range, $\triangle C$ is the amount of capacitance added to or subtracted from the capacitance across the emitter resistance. The 10-per-cent to 90-per-cent rise time T_r is then given by

$$T_r = 2{\cdot}2\,T_e \qquad (30)$$

The following example illustrates the design procedure.

Problem

A pulse amplifier similar to that shown in Fig. 15 is required to have the following characteristics : voltage gain A=20, input impedance R_{in}= 75-ohm cable, load capacitance C_{load}=10 picofarads, supply voltage V_{cc}=28 volts, maximum temperature T_{max}=75 degrees Centigrade.

Calculation

The free-air dissipation P_c of the 2N1491 is 0·25 watt at an ambient temperature of 75 degrees Centigrade. Therefore,

$$I_c = \frac{P_c}{\dfrac{V_{cc}}{2}} = \frac{(0{\cdot}25)}{\dfrac{28}{2}} = 17{\cdot}8 \text{ ma}$$

$$R_c = \frac{\dfrac{V_{cc}}{2}}{I_c} = 785 \text{ ohms}$$

(Use 750 ohms)

The gain of the amplifier A is given by

$$A = \frac{g_m\,R_L\,R_{b'e}}{R_{b'e}\,(1+g_m\,R_e)+r_{bb'}}$$

The given values are :

$$A = 20$$
$$r_{bb'} = 30 \text{ ohms}$$
$$R_{b'c} = 60 \text{ ohms}$$
$$g_m = 0{\cdot}38 \quad I_e = (0{\cdot}038)\,(17{\cdot}8)$$
$$= 0{\cdot}675 \text{ mho}$$

Substitution of these values in the gain equation yields a value of approximately 33 ohms for R_e.

$$C_i = C_{b'e}+C_{b'c}\,g_m\,R_L$$
$$= 450+2(0{\cdot}675)\,(750)$$
$$= 1462 \text{ pf}$$
$$R_i = R_{b'e}\,(1+g_m\,R_e)$$
$$= 60\,[1+(0{\cdot}675)\,(33)]$$
$$= 1396 \text{ ohms}$$

$$R_{in} = \frac{R_i\,(r_{bb}' + R_g)}{R_i + r_{bb}' + R_g}$$

$$= \frac{(1396)\,[(30)+(75)]}{1396+30+75}$$

$$= 97{\cdot}6 \text{ ohms}$$

$$T_{in} = R_{in}\,C_{in}$$

$$= \frac{R_{in}\,C_i}{1+ g_m\,R_e}$$

$$= 6{\cdot}15 \text{ nanoseconds}$$

The capacitive loading on the output C_o is given by

$$C_o = C_{ob} + C_{wiring} + C_{load}$$

$$= 5 + 3 + 10$$

$$= 18 \text{ pf}$$

Assume, for trial, that the $R_e\,C_e$ product is equal to the $R_L\,C_L$ product. Then the output time constant T_{out}, with shunt peaking, is given by

$$T_{out} = \frac{R_L\,C_o}{1{\cdot}7(1+g_m\,R_e)}$$

$$= \frac{(75)\,(18)\,(10^{-12})}{1{\cdot}7[1+(0{\cdot}675)\,(33)]}$$

$$= 0{\cdot}61 \text{ nanosecond}$$

Note how much lower the output time constant is than the input time constant T_{in}. The response of the amplifier is, therefore, determined primarily by the input network. The value of C_e may be determined as follows :

$$R_e\,C_e = (T_{in}^2 + T_{out}^2)^{\frac{1}{2}} = T_{in}$$

$$C_e \cong \frac{T_{in}}{R_e} = \frac{6{\cdot}15\times10^{-9}}{33}$$

$$= 185 \text{ pf}$$

The upper cutoff frequency is then given by

$$f_{3db} = \frac{1}{2\pi\, 6{\cdot}15\times 10^{-9}}$$

$$= 25{\cdot}9 \text{ megacycles}$$

The rise time T_r is given by

$$T_r = 2{\cdot}2(6{\cdot}15)$$

$$= 13{\cdot}5 \text{ nanoseconds}$$

The gain-bandwidth product of the amplifier is $25{\cdot}9\times20$, or 518. This product remains approximately the same, at this emitter current, for small changes in R_e or R_L, and may be used as a guide for the design of amplifiers requiring a different gain or bandwidth from that shown in the example.

The final step is to obtain the values of the biasing resistors R_1, R_2, and R_3. The collector-to-emitter voltage drop may be assumed to be equal to the drop in the load resistance, minus one volt. The resistor values are then determined as follows :

$$R_1 + R_3 = \frac{V_{cc} - 2(R_L)\,(I_c) - 1}{I_c}$$

$$= \frac{28 - 2(750)\,(17{\cdot}8\times10^{-3}) - 1}{17{\cdot}8}$$

$$R_1 = 129 - 33 = 96 \text{ ohms}$$

(use 100 ohms)

If it is assumed that very stable operation is required, an S factor of 5 should be included to cover the most extreme environments.

$$R_3 = \frac{V_{cc}(S-1)}{I_c - S_{1co}}$$

$$= \frac{28(4)}{17{\cdot}8 - 5(0{\cdot}01)}$$

$$= 6280 \text{ ohms (use 6200 ohms)}$$

$$R_2 = \frac{(R_1+R_e)\,R_3(S-1)}{R_3\,S\alpha_{fb} - (S-1)\,(R_1+R_3)\,(R_e)}$$

$$= \frac{(133)\,(6200)\,(4)}{6200\,(5)\,(0{\cdot}95) - (5-1)\,(6333)}$$

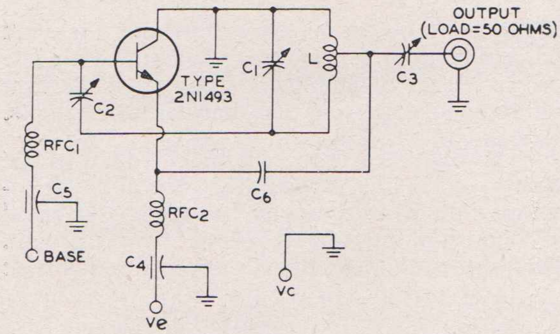
$$= 810 \text{ ohms (use 820 ohms)}$$

Typical Circuits

The typical circuits shown below have been designed for the 2N1491, 2N1492, and 2N1493, using the design equations and characteristics curves discussed previously. These circuits have actually been constructed and tested, and have been in operation in equipment for some time.

Oscillators

Three 70-megacycle power oscillators are shown in Figs. 16, 17, and 18. Although the 2N1491 is shown in these circuits, either the 2N1492 or the 2N1493 could be used with the proper collector voltage to provide higher power output. In the circuit of Fig. 16, an output of 0·6 watt is obtained when a typical 2N1493 is used at a collector voltage of 50 volts and a collector current of 25 milliamperes. The case of the transistor is clamped down firmly to the copper chassis to provide a low-thermal-resistance path to the

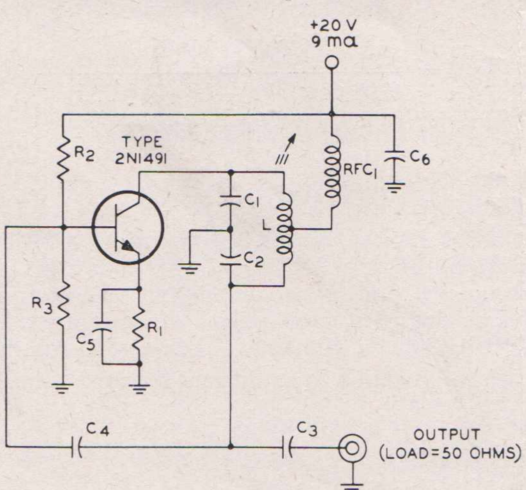


C_1: 5-50 $\mu\mu f$, variable
C_2, C_3: 7-100 $\mu\mu f$, variable
C_4, C_5: feedthrough capacitor, 1500 $\mu\mu f$
C_6: 0.01 μf
L: 3-1/2 turns No. 14 wire, 3/4-inch diameter, tapped at 2 turns
RFC_1, RFC_2: radio-frequency choke, 10 μh

Fig. 16—70-megacycle power oscillator.

chassis. No insulating mica spacers are required because the collector is also connected to the chassis.

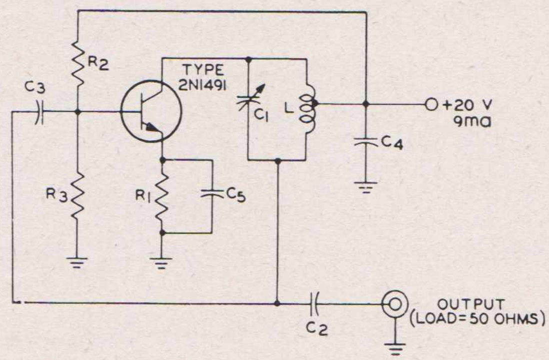
Low-power vhf oscillators are required in many applications, including local oscillators and variable-frequency oscillators in transmitters. In many of these applications, an output power of about 25 or 50 milliwatts is sufficient. The Colpitts oscillator shown in Fig. 17 has a maximum



R_1: 200 ohms
R_2: 9100 ohms
R_3: 1500 ohms
C_1: 20 $\mu\mu f$
C_2: 30 $\mu\mu f$
C_3: 22 $\mu\mu f$
C_4: 18 $\mu\mu f$
C_5, C_6: 0.01 μf
L: 5 turns No. 20 wire close wound on CTC LS5 form, tapped at 3 turns (white dot core)
RFC_1: radio-frequency choke, 10 μh

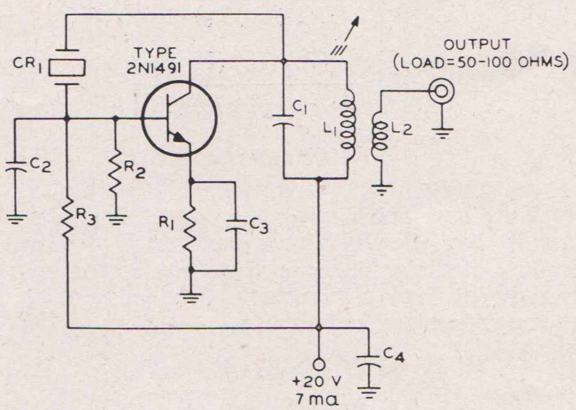
Fig. 17—70-megacycle Colpitts oscillator.

output of up to 50 milliwatts and a typical value of 35 milliwatts. The Hartley oscillator shown in Fig. 18 has approximately the same output capabilities. The 2N1492 or 2N1493 may also be used in these circuits if higher power output is required.



R_1: 200 ohms
R_2: 9100 ohms
R_3: 1500 ohms
C_1: 4-24 $\mu\mu f$, variable
C_2: 12 $\mu\mu f$
C_3: 15 $\mu\mu f$
C_4, C_5: 0.01 μf
L: 3 turns No. 18 wire, 3/4-inch diameter, tapped at 2 turns

Fig. 18—70-megacycle Hartley oscillator.



R_1: 200 ohms
R_2: 680 ohms
R_3: 9100 ohms
C_1: 22 $\mu\mu f$
C_2: 20 $\mu\mu f$
C_3, C_4: 0.01 $\mu\mu f$
CR_1: 27 Mc crystal
L_1: 15 turns No. 22 wire close wound on CTC LS5 form (white dot core)
L_2: 2 turns No. 18 wire wound over cold end of L_1

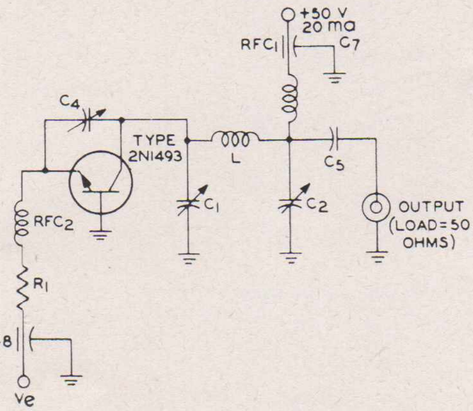
Fig. 19—27-megacycle crystal oscillator.

Many applications require crystal-controlled oscillators to provide a sufficiently stable output frequency. Fig. 19 shows a crystal-controlled oscillator which has good frequency stability and uses overtone crystals in the frequency range of 25 to 30 megacycles. The circuit, using a typical 2N1491, is capable of a power output of 4 milliwatts into a load resistance of 50 to 100 ohms.

Fig. 20 shows a 250-megacycle oscillator circuit using a typical 2N1493. This circuit provides an output of 150 milliwatts at 250 megacycles with an efficiency of from 15 to 20 per cent. The output matching circuit, a pi type, reduces the harmonics in the load. In this circuit, as in all other circuits operating at high frequency, it is very important to keep lead lengths short and minimize stray capacitance.

Power Amplifiers

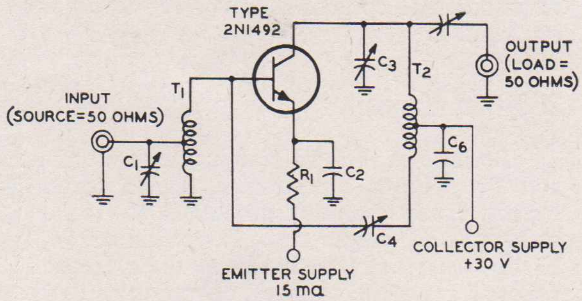
Fig. 21 shows a 70-megacycle power amplifier which may be used in intermediate or final power-amplifier circuits to amplify the output of a low-level oscillator or a crystal oscillator. The 2N1492 provides a power output of 100 milliwatts with a power gain of 15 db at a collector voltage of 30 volts and a collector current of 15 milliamperes. If a 2N1493 is used, a power output of one-half watt with a typical power gain of 12 db is obtained.



C_1: 3-15 $\mu\mu$f
C_2: 4-50 $\mu\mu$f
C_4: 0.6-5.5 $\mu\mu$f
C_5: 0.002 μf
C_7, C_8: feedthrough capacitor, 1000 $\mu\mu$f
R_1: 400 ohms
RFC_1, RFC_2: radio-frequency choke, 0.82 μh
L: 1 turn No. 14 wire, 1-inch diameter

Fig. 20—250-megacycle oscillator.

The 50-megacycle amplifier circuit shown in Fig. 22 is a high-efficiency high-power-output amplifier. It is designed for use in portable or mobile transmitters as a driver or output stage. This circuit provides a power output of one-half watt with a power gain of 10 db at a collector voltage of 28 volts and a collector current of 23 milliamperes. At a higher collector voltage of 50 volts and a collector current of 20 milliamperes an output of one watt can be obtained with a power gain of 8 db.



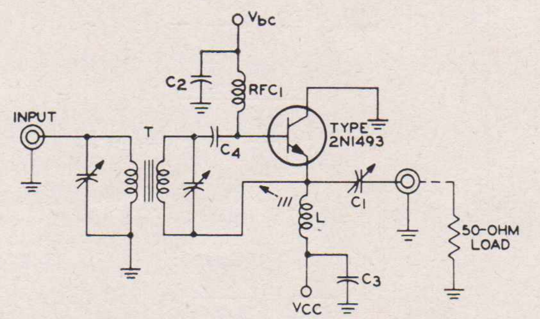
C_1: 3-20 $\mu\mu$f, variable
C_2, C_6: 0.01 μf
C_3: 3-20 $\mu\mu$f, variable
C_5: 3-20 $\mu\mu$f, variable
R_1: 1000 ohms, 2 watts
T_1: 8 turns No. 24 wire on General Ceramics Corp. F303 toroid of Q material tapped at 1 turn
T_2: 8 turns No. 24 wire wound on CTC 3/8-inch ceramic coil form (no slug used), tapped at 2.5 turns

Fig. 21—70-megacycle power amplifier.

The circuit performs as a common-emitter amplifier even though the collector is grounded because both the emitter and the base are isolated from ground. The potential of the entire secondary winding of the transformer varies with the load voltage, whereas the potential difference across the secondary winding drives the transistor. The design of the output network is similar to that shown for the 250-megacycle amplifier discussed later.

The amplifier of Fig. 22 is biased deep into class C ; the magnitude of the bias depends on the driving power available. Increased drive permits more cutoff bias and produces higher efficiency.

In operation, the transistor case should be firmly clamped to a metal chassis. This arrangement provides an adequate built-in heat sink. A 3-inch by 3-inch chassis is suitable for most



C_1: 3-15 $\mu\mu$f, variable
C_2, C_3: 0.01 μf
C_4: 500 $\mu\mu$f
RFC_1: choke, 4.7 μh
L: variable inductor, 0.6-0.8 μh
T: Toroidal transformer with minimum capacitance between windings

Fig. 22—50-megacycle amplifier.

environmental conditions at power levels up to one watt.

250-Mc Amplifier

Because of their high alpha-cutoff frequencies and low output capacitances, RCA silicon vhf transistors are excellent amplifiers at frequencies as high as 300 megacycles. Power gains from 8 to 13 db are readily obtained at these frequencies depending on the class of operation and power output desired.

A single-tuned 250-megacycle amplifier is shown in Fig. 23. At typical operating conditions, this amplifier will produce 13 db of power gain at an output level of 10 milliwatts. The amplifier is quite stable and operates over a wide temperature range.

Because the input is inductive, a single capacitor may be used to tune the input network at a 250-megacycle frequency. The slight resistive mismatch is not significant.

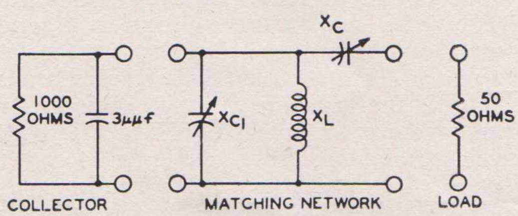


Fig. 23—L matching network used in output section of 250-megacycle amplifier.

An L matching network such as that shown in Fig. 24 is used in the output section because of its simplicity and efficiency. The design procedure for this network is given below :

$$Q = \sqrt{\frac{R_o}{R_L} - 1}$$

$$= \sqrt{\frac{1000}{50} - 1'} = 4{\cdot}35$$

$$X_C' = Q\,R_L = (4{\cdot}35)\,(50) = 218 \text{ ohms}$$

$$X_L' = \frac{R_o}{Q} = \frac{100}{4{\cdot}35} = 230 \text{ ohms}$$

It may be assumed that the total of the output capacitance C_{ob} and the tuning capacitance is 8 picofarads. At 250 megacycles, this capacitance is equivalent to a reactance of approximately 80 ohms. X_L can then be calculated as follows :

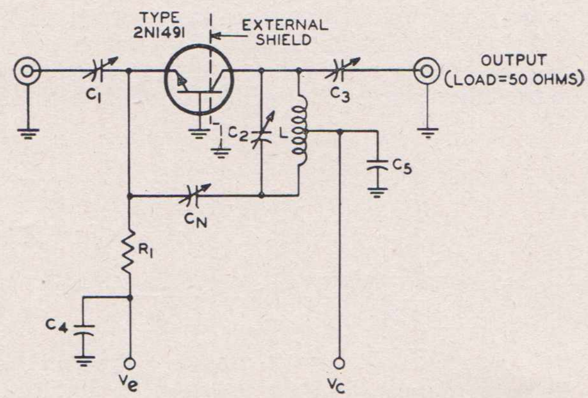
$$X_L = \frac{X_L'\,X_c}{X_L' + X_c}$$

$$= \frac{(218)\,(80)}{218+80} = 58{\cdot}5 \text{ ohms}$$

This value is equivalent to an inductance of approximately 0·035 microhenry.

$$C_1 = C_t - C_{ob} = 5 \text{ pf}$$

If a balanced-bridge network is used for neutralization, X_L and X_{c1} remain as calculated when the output inductance is tapped at its centre. The neutralization capacitor is then approximately equal to $C_{b'c}$, which is between 2 and 3 pf.



C_1: 5-40 μμf, variable
C_2, C_3, C_N: 0.8-8 μμf, variable
C_4, C_5: 0.001 μf
R_1: 820 ohms
L: 0.035 μh, center tapped

Fig. 24—250-megacycle amplifier.

High-Input-Impedance Video Amplifier

One of the disadvantages of transistors, when compared to vacuum tubes, is their low input impedance. It is possible, however, to raise the equivalent input impedance of a transistor to any value required by the driving source by the use of appropriate circuits.

Equivalent input impedance can be increased by the use of negative feedback or a combination of negative and positive feedback, as shown in Figs. 25 and 26. These configurations have the effect of trading gain for input impedance.

Fig. 25 shows a circuit which employs only negative feedback. Resistor R_{FB} provides the path for a portion of the output voltage to be returned to the emitter of the first stage. This voltage, in phase with the input signal, reduces the net voltage drop between emitter and base ; input loading is reduced correspondingly.

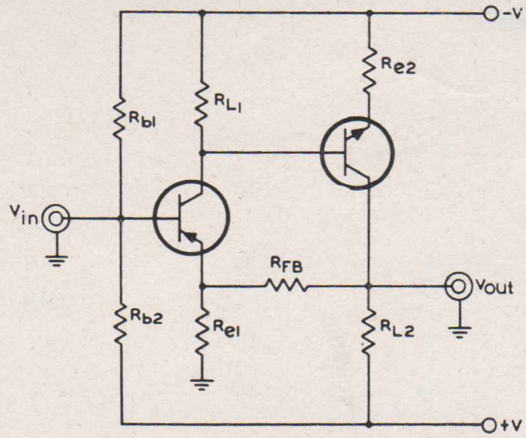


Fig. 25—Video amplifier having negative feedback.

If one transistor is p-n-p and the other n-p-n, larger bias resistors R_{b1} and R_{b2} may be used to reduce the loading produced by the bias network.

The gains of the first and second stages, A_1 and A_2, are given by

$$A_1 = \frac{g_{m1}\, R_{L1}}{1+g_{m1}\, R_{e1}}$$

$$A_2 = \frac{g_{m2}\, R_{L2}}{1+g_{m2}\, R_{e2}}$$

The feedback network provides a gain reduction factor κ given by

$$\kappa = \frac{R_{e1}}{R_{e1}+R_{FB}}$$

The closed-loop gain A_{12} may then be calculated as follows :

$$A_{12} = \frac{A_1\, A_2}{1+\kappa\, A_1 A_2}$$

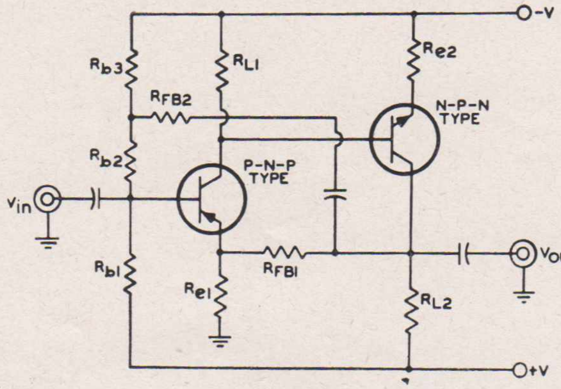


Fig. 26—Video amplifier having negative and positive feedback.

A positive-feedback loop may be provided to offset the loading effects of the bias resistors, as shown in Fig. 26. Resistor R_{FB2} is adjusted to produce a voltage at the junction of R_{b2} and R_{b3} which is in phase with and equal to the input voltage. When the voltage drop across the bias resistors is reduced, the result is equivalent to raising the resistance of the network.

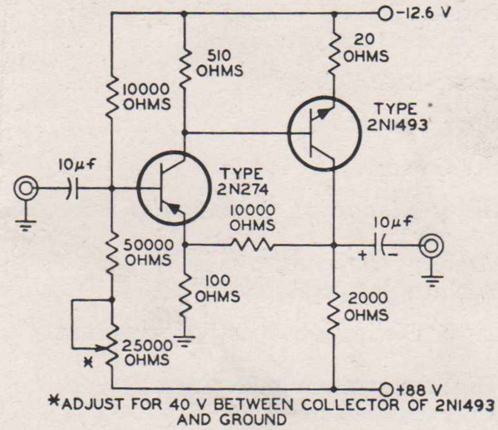


Fig. 27—High-input-impedance video amplifier.

A typical video-amplifier circuit employing negative feedback is shown in Fig. 27. The specifications of the circuit are as follows : frequency response=20 cycles to 7·5 megacycles (15—pf load) ; voltage gain=75 ; input resistance=5000 ohms. This circuit may be used in industrial-television video amplifiers, instrument amplifiers, or pulse amplifiers.

(With acknowledgements to RCA)

Technical Data on transistors types 2N1491, 2N1492 and 2N1492 used as examples in this article will be found overleaf.

Transistor Data

Maximum Ratings, *Absolute-Maximum Values :*

	2N1491	2N1492	2N1493	
Collector-to-Base Voltage	30	60	100	volts
Collector-to-Emitter Voltage (with emitter to base reverse biased)	30	60	100	volts
Emitter-to-Base Voltage	1	2	4·5	volts
Collector Current	50	50	50	ma
Emitter Current	50	50	50	ma
Transistor Dissipation Operation in free air :				
Ambient temperature of 25 °C. ..	0·5	0·5	0·5	watt
Ambient temperature of 100 °C. ..	0·25	0·25	0·25	watt
Transistor Dissipation Operation with heat sink :				
Case temperature of 25 °C.	3	3	3	watts
Case temperature of 100 °C.	1·5	1·5	1·5	watts
Ambient Temperature Range :				
Operating and storage	−65 to +175	−65 to +175	−65 to +175	°C.

Characteristics

	2N1491	2N1492	2N1493	
Minimum Collector Breakdown Voltage (at $I_c=0{\cdot}1$ ma ; $I_e=0$)	30	60	100	volts
Maximum Collector Cutoff Current (at $V_{CB}=12$ volts ; $I_e=0$)	10	10	10	μa
Maximum Emitter Cutoff Current (at $V_{EB}=0{\cdot}5$ volt ; $I_c=0$)	100	100	100	μa
Maximum Collector-to-Base and Stem Capacitance* : (at $V_{CB}=30$ volts ; $I_e=0$)	5	5	5	pf
Alpha-Cutoff Frequency* : (at $V_{CB}=30$ volts ; $I_c=15$ ma)	250	275	300	Mc
Small-Signal Current Transfer Ratio** :				
(at 1Kc ; $V_{CE}=20$ volts ; $I_c=15$ ma)	50	50	50	
(at 100 Mc ; $V_{CE}=30$ volts ; $I_c=15$ ma)	1·8	1·8	1·8	
Power Gain at 70 Mc** :				
(Power output=10 mw ; $V_{CB}=20$ volts ; $I_e=-15$ ma)	15	—	16	db
(Power output=100 mw ; $V_{CB}=30$ volts ; $I_e=-15$ ma)	—	15	16	db
(Power output=500 mw ; $V_{CB}=50$ volts ; $I_e=-25$ ma)	—	—	12	db
Maximum Thermal Resistance :				
Junction-to-case	50	50	50	°C/watt

* Common-Base Circuit, Emitter Input

** Common-Emitter Circuit, Base Input

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