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SHOCK EXCITATION IN RADAR RECEIVER CIRCUITS

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ABSTRACT

Electrical shock excitation of the input circuits of the intermediate-frequency amplifier of a radar system is discussed. The deleterious effects of shock from the trailing edge of a strong pulse are discussed. Fidelity of reproduction of the trailing edge of a high-amplitude pulse is shown to be related to the frequency at which the amplifier input circuit is resonant. The pulse fall time is shown to be an important factor in the choice of an intermediate frequency. Although the report is written with reference to a marine navigational radar having high definition at short range, it may be of value in the design of other radar systems where shock effects following a high-amplitude pulse cannot be tolerated.

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SHOCK EXCITATION IN RADAR RECEIVER CIRCUITS

- W.M. Cameron -

INTRODUCTION

Two basic requirements for a minimum-range marine navigational radar are: use of a short transmitter pulse, which may have a length of 10^{-1} microsecond or less, and provision of sufficient receiver bandwidth to pass the pulse with some permissible amount of distortion. It is believed that a satisfactory radar display can be secured if the intermediate-frequency bandwidth in megacycles per second is 1.5 times the reciprocal of the pulse length in microseconds. A third requirement is the use of a suitable intermediate frequency.

Best close-in definition can be obtained only if the region immediately following the transmitter pulse is free from spurious signals. Offending signals may be induced into the receiver by the modulator, which in many designs is adjacent to the input section of the intermediate-frequency amplifier. Judicious placement of modulator components, adequate shielding, and filtering of the power leads at their point of entry into the intermediate-frequency amplifier will do much to reduce interference from this source. Ground loops must be avoided.

On the other hand, the intermediate-frequency amplifier may be shock-excited directly by the pulse from the mixer crystal. Preliminary observations indicated that, in the case of one radar, superior minimum-range performance resulted if the standard 30-Mc/s intermediate-frequency amplifier were replaced by a 60-Mc/s strip.

INSTRUMENTATION

Experimental amplifiers were built for frequencies of 30 Mc/s, 60 Mc/s, and 120 Mc/s, to permit a critical study of circuit shock excitation by the radio-frequency pulse. The same circuit (see Fig. 1) was used in each of the three amplifiers, with the exception of the values of the band-pass elements. In all cases the bandwidth of the interstage circuit was 10 Mc/s. The output circuit was adjusted to a Q of 0.5, and hence could not be shock-excited because of the critical damping. The input circuits of an intermediate-frequency amplifier should be designed for best noise figure, but for the bandwidth required in minimum range radar systems, this becomes impossible at 30 Mc/s. Therefore, a transitionally coupled circuit of sufficient bandwidth, without regard to noise figure, was used in this amplifier. The design Q's of the input circuits used in the three amplifiers were very nearly the same.

RELATIVE AMPLIFIER RESPONSE MEASUREMENTS

The impedance design center of the amplifier input circuits was taken as

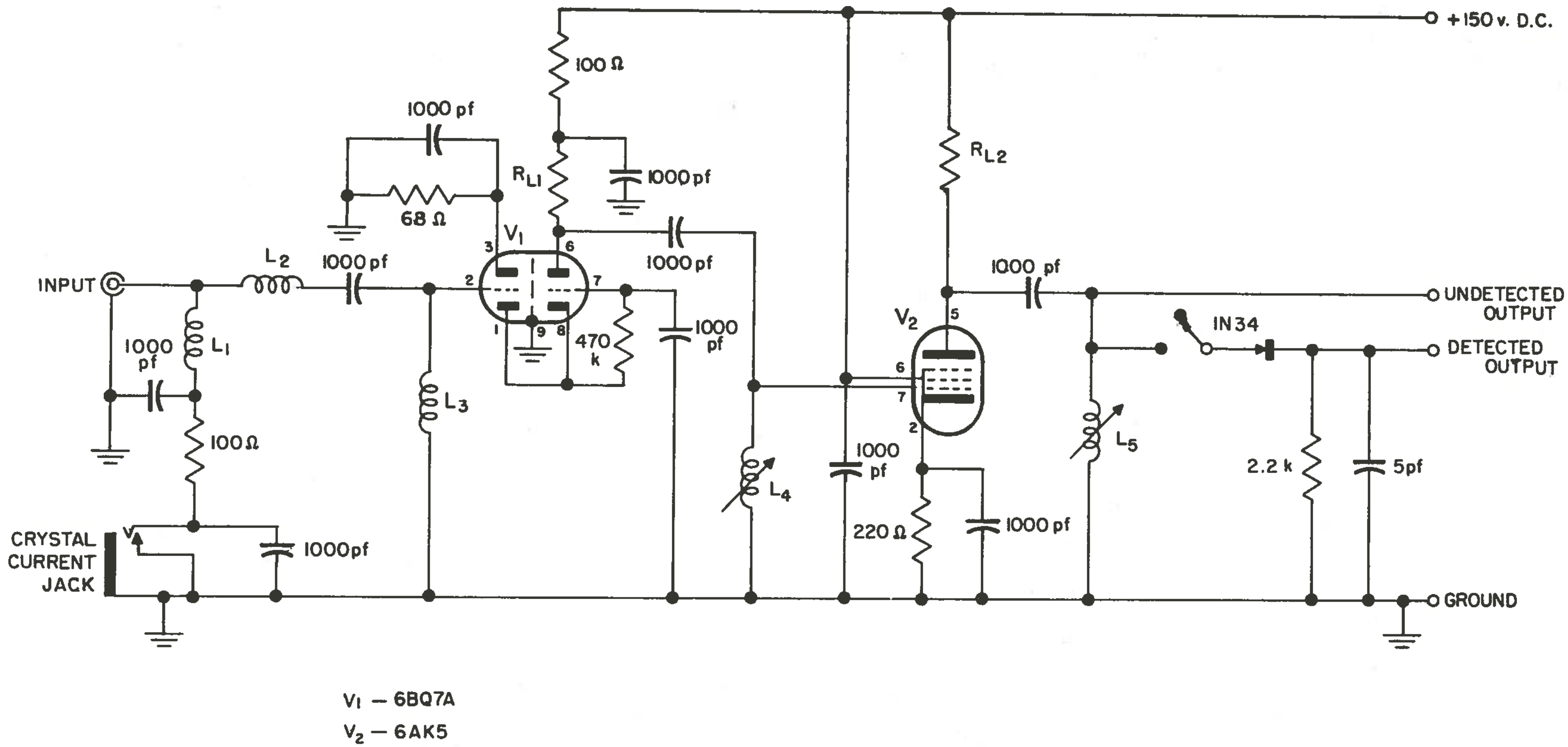


FIG. 1. CIRCUIT DIAGRAM OF TEST AMPLIFIERS

380 ohms, and it was necessary to insert a resistive L-pad in the signal generator lead to obtain this value. The relative division of signal energy between the L-pad and the input circuit could not be established as, in general, the input circuit is not necessarily a match for this impedance. Hence relative gain measurements could not be made using a signal generator having the usual 50-ohm output impedance.

It was decided to measure relative pulse gains of the amplifiers when operating in a radar set. The radio-frequency pulse produced by the magnetron was attenuated and taken through waveguide to a crystal mixer. Sufficient attenuation was used to reduce the output of the reference amplifier to 0.4 volt. The same mixer crystal was used throughout the measurements and the crystal current was maintained at 0.5 milliampere. Local oscillator tuning was adjusted for the maximum amplitude of a flat pulse in each measurement. The pulse length was 0.18 μsec, as shown in Fig. 2 (a).

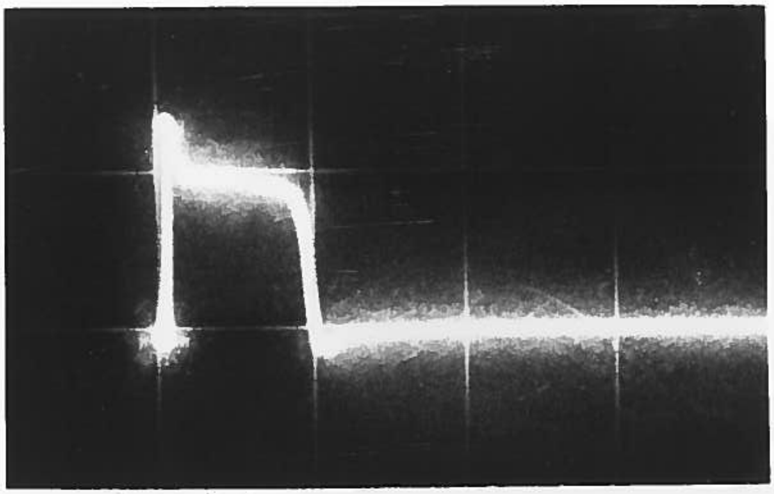
The relative gains of the amplifiers were measured with reference to that of the 60-Mc/s amplifier, using a calibrated attenuator in the waveguide. The gain of the 30-Mc/s amplifier was found to be +8 db, and that of the 120-Mc/s amplifier -10 db, with respect to the gain of the 60-Mc/s amplifier. In each case, 6 db can be accounted for in the critically damped stage as, for the same stage capacitance, the required value of damping resistor is inversely proportional to frequency. The residual terms, +2 db and -4 db, were taken to be real, and are believed to be due to increased losses at the higher frequencies.

The amplifier to be tested was substituted for the intermediate-frequency amplifier in a typical common T-R radar system. The radar waveguide was terminated in a flat load. The output of the amplifier was displayed on a high-speed oscilloscope having a distributed-line amplifier. Since normalization over an 18-db range was required, an additional distributed-line amplifier and calibrated attenuator was inserted between the test amplifier output and the oscilloscope input probe. All connector lengths were kept to a minimum to reduce direct pickup from the modulator. However, some modulator interference was experienced. The spurious signal at 120 Mc/s with no attenuation is shown in Fig. 3 (a); the 30-Mc/s pickup attenuated 18 db to compensate for the higher gain of the amplifier appears in Fig. 3 (b). These oscillograms were taken with the waveguide blocked, and are almost certainly the result of direct pickup of modulator transients by circuits external to the amplifier.

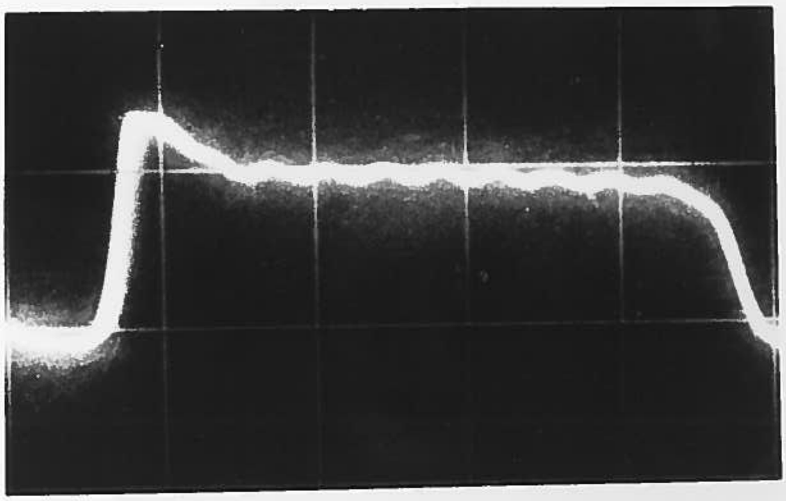
RESULTS

AMPLIFIER SHOCK EXCITATION

Two oscillograms were taken to identify the circuit of the 30-Mc/s amplifier

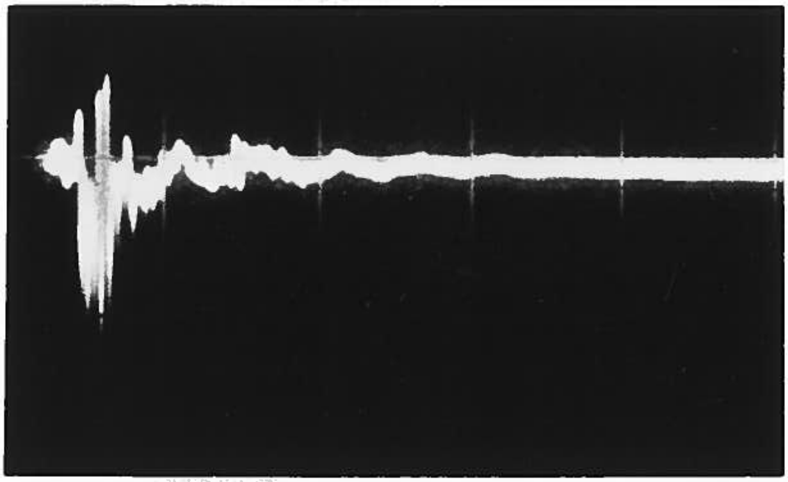


(a) Sweep speed 0.2 μsec per division

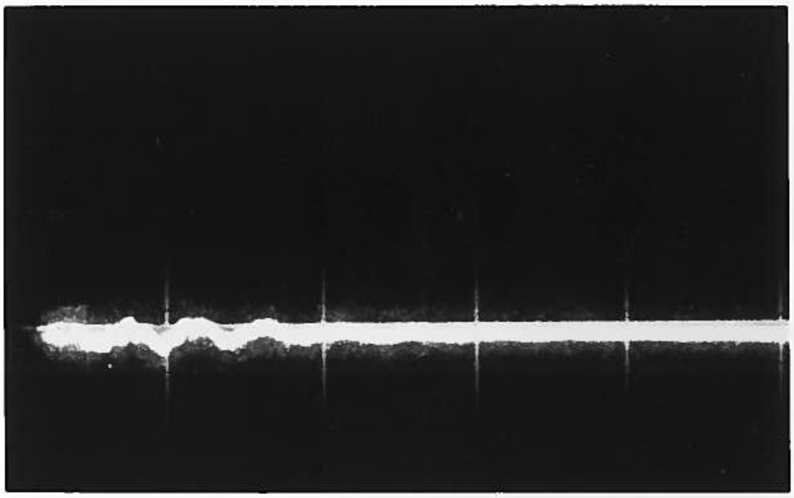


(b) Sweep speed 0.05 μsec per division

FIG. 2 RECTIFIED TRANSMITTER PULSE FROM MIXER CRYSTAL WITH 330-OHM TERMINATION



(a) No attenuation



(b) Attenuation = 18 db

FIG. 3 MODULATOR INTERFERENCE
(Waveguide to crystal mixer blocked;
sweep speed 0.2 μsec per division)

being shock-excited, as the relative contributions of the input and interstage circuits were not known. The detector circuit in the amplifier was disconnected and the oscilloscope input probe connected to the "undetected output" terminal (Fig. 1) to obtain the oscillograms shown in Figs. 4 (a) and 4 (b). Fig. 4 (a) shows the shock excitation which occurs with the local oscillator inoperative. Note that both leading and trailing edges of the pulse initiate a wave train. For convenience, the former will be referred to as primary shock, and the latter as secondary shock. Analysis of the latter part of the wave in Fig. 4 (a) indicates a Q of about 15. As the interstage circuit had a Q of 3, and hence could not offend in the region under consideration, it would seem that this portion of the wave must arise elsewhere.

Fig. 4 (b) was taken under the same conditions, except that the local oscillator was tuned. The secondary-shock wave train indicates a lower Q than in Fig. 4 (a), and this can be attributed only to a damping action because of local oscillator injection. Also, it was noted that, as the local oscillator tuning approached optimum, the apparent Q of the offending circuit decreased.

From the above, it may be concluded that the wave train arises from the input circuit, and that the increased damping shown in Fig. 4 (b) occurs because the mixer crystal is rendered conductive by local oscillator injection. The contribution of successive amplifier stages at bandwidths under consideration appears to be much less significant.

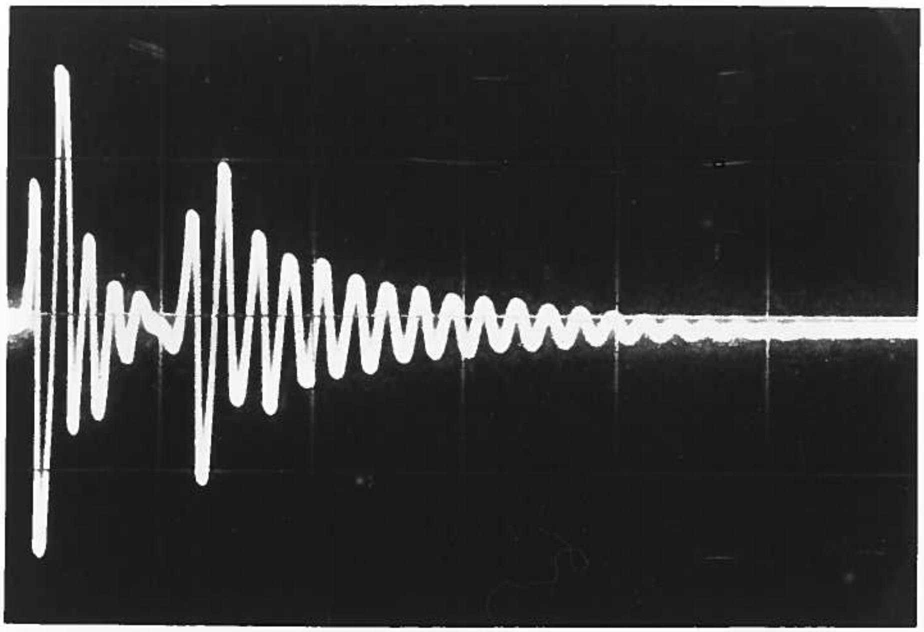
RELATIVE SUSCEPTIBILITY TO SHOCK

The oscillograms shown in Figs. 5 (a), (b), and (c) apply to the 30-Mc/s, 60-Mc/s, and 120-Mc/s amplifiers, respectively, with no local oscillator injection. The oscilloscope probe was connected to the detector output of the amplifiers, as the oscilloscope amplifier was not suitable for use at the higher frequencies. The system gain was equalized in each case.

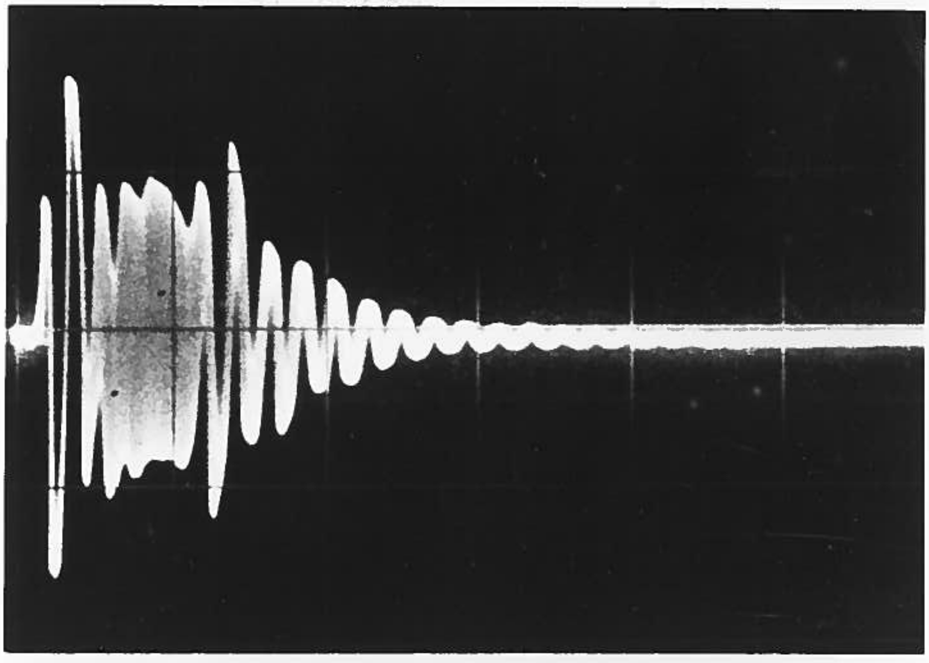
A significant improvement over the performance shown in Fig. 5 (a) is apparent in Fig. 5 (b), where the secondary shock effect is very small.

A true evaluation of shock in the 120-Mc/s amplifier may be obtained by subtracting the stray pickup shown in Fig. 3 (a) from the waveform in Fig. 5 (c). It will be seen that secondary shock is undetectable.

Figs. 6 (a), (b), and (c) show the response of the 30-Mc/s, 60-Mc/s, and 120-Mc/s amplifiers, respectively, to the 0.18 μsec pulse. These oscillograms were taken from the detector output of the test amplifiers and the local oscillator was tuned for best pulse shape and maximum pulse amplitude. The mixer crystal current was held at 0.5 milliampere.

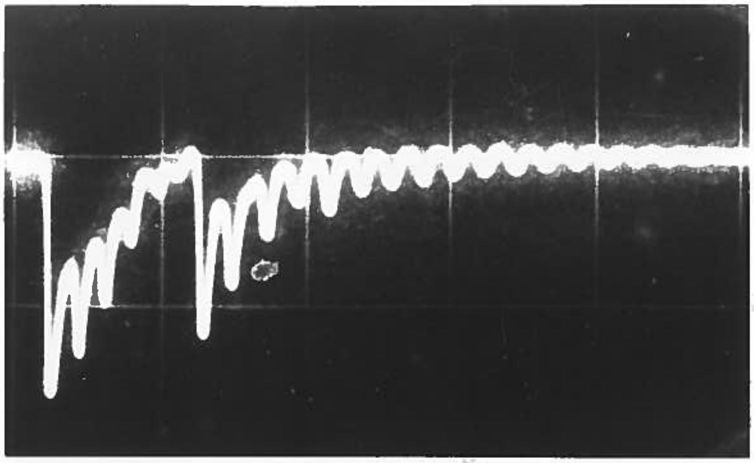


(a) Local oscillator not functioning

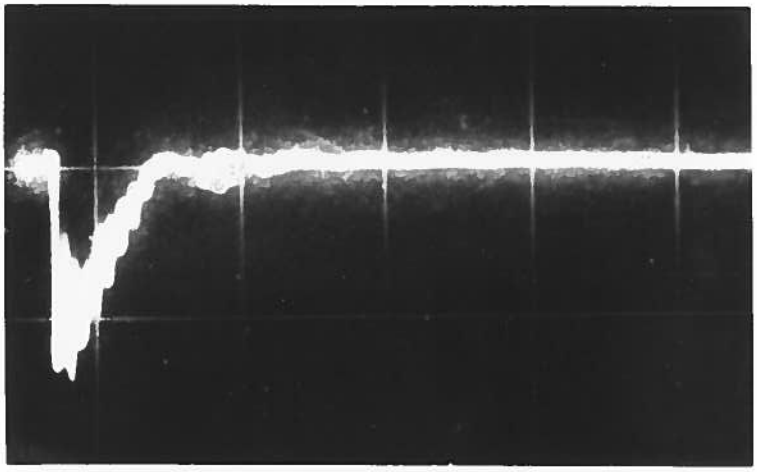


(b) Local oscillator tuned for best pulse shape and minimum secondary shock

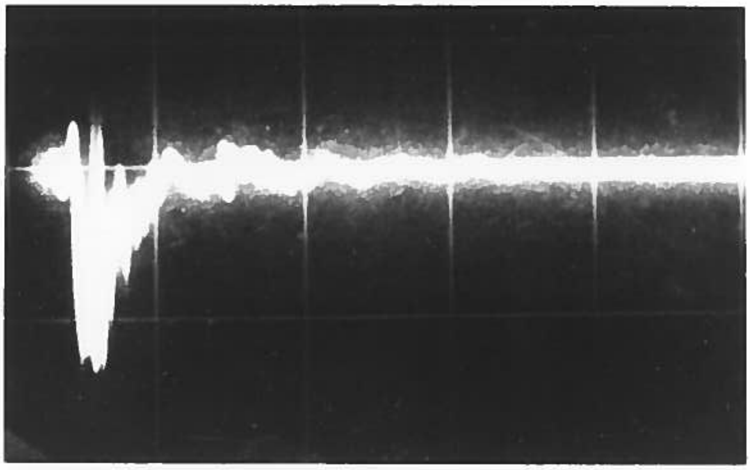
FIG. 4 INTERMEDIATE-FREQUENCY RESPONSE OF 30-MC/S AMPLIFIER TO PULSE FROM MIXER CRYSTAL — USED IN DETERMINATION OF CIRCUIT Q (sweep speed 0.2 μsec per division)



(a) 30-Mc/s Amplifier

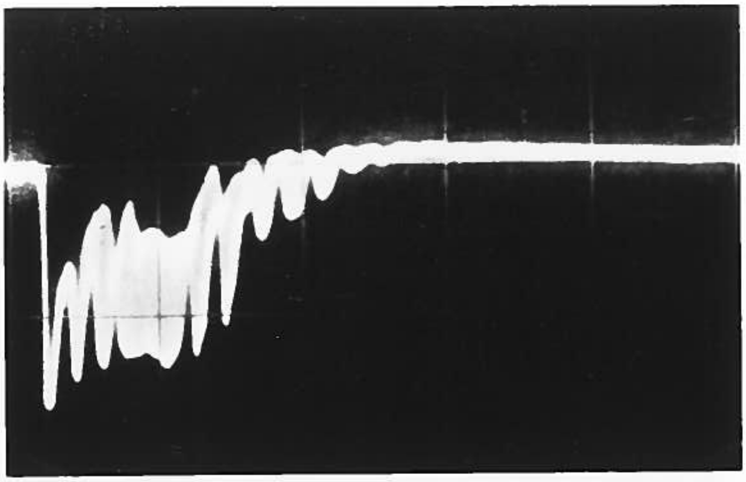


(b) 60-Mc/s Amplifier

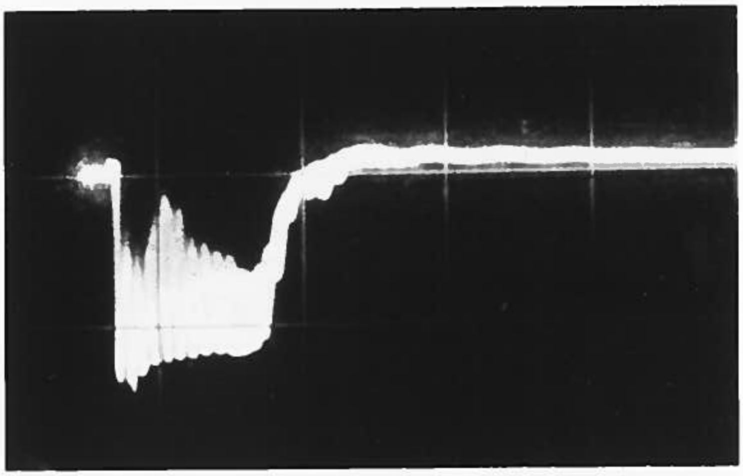


(c) 120-Mc/s Amplifier

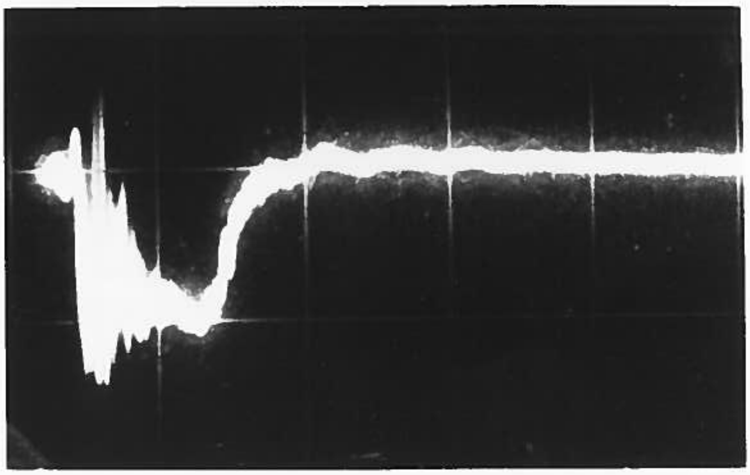
FIG. 5 SHOCK EFFECTS AFTER DETECTION
(Local oscillator not functioning — sweep speed 0.2 μsec per division)



(a) 30-Mc/s Amplifier



(b) 60-Mc/s Amplifier



(c) 120-Mc/s Amplifier

FIG. 6 REPRODUCED TRANSMITTER PULSE AFTER DETECTION
(Local oscillator tuned — sweep speed 0.2 μsec per division)

CONCLUSION

One must consider the effects of further amplification, amounting to 80 or 90 db, to obtain an appreciation of the final result, viz., a radar PPI picture. Because of the asymptotic nature of the exponential decay, additional amplification will show shock effects extending over a greater period of time than that evident in the oscillograms. Any signal which is detectable in Figs. 6 (a), (b), and (c) will almost certainly appear as a saturation signal in a conventional radar receiver at full gain. A decrease of receiver gain for viewing close-in echoes is normal radar operating practice, but sufficient gain must be retained for echo registration. This would probably be of the order of 30 to 40 db above equalized test amplifier gain.

CONDITIONS FOR MINIMUM RANGE

In this discussion of minimum range, the following conditions are presumed to have been satisfied.

a) Separate antennas are used for transmission and reception.
b) Excellent isolation exists between modulator and receiving system.
c) Receiver gain is set at 40 db above that at which the oscillograms of Figs. 6 (a), (b), (c) were taken. (Adequate video bandwidth is, of course, assumed.)
d) A transmitter pulse length of 0.18 μsec and a receiver bandwidth of 10 Mc/s are used.
e) Minimum-range echoes are presumed to arise from small close-in objects, as, for example, spar buoys.
f) Sea clutter is assumed to be absent.

Relative minimum range can then be estimated in terms of the masking effect of spurious signal immediately following the transmitter pulse. This effect will appear in the reproduced transmitter pulse because of secondary shock excitation. The masking effect is defined as $\tau_2 - \tau_1$, where τ_2 is the reproduced transmitter pulse length and τ_1 is the actual transmitter pulse length.

In the interpretation of the oscillograms, Figs. 6 (a), (b), (c), and 2 (a), pulse lengths should be measured at points where the amplitude is 40 db down to take account of the additional postulated 40 db of receiver gain. Such a measurement would correspond to about 0.2 millimeter on the scale of the oscillograms and, of course, is impractical. Hence, reference is taken as zero amplitude. This gives a value of 0.2 μsec for τ_1. The relative performance at the three intermediate frequencies is shown in Table I.

TABLE I

MASKING EFFECT AS A FUNCTION OF INTERMEDIATE FREQUENCY

f (Mc/s)	τ_2 (μsec)	$\tau_2 - \tau_1$ (μsec)	$\tau_2 - \tau_1$ (equivalent range in yards)
30	0.45	0.25	42.0
60	0.27	0.07	11.7
120	0.25	0.05	8.4

In practice the masking effect at 120 Mc/s could be made even smaller by an increase of amplifier bandwidth and a reduction of detector time constant.

APPENDIX

EFFECT OF CIRCUIT Q

The Q of a circuit is given by $\frac{f_0}{f_2 - f_1}$, where f_0 is the midband frequency, and f_2 and f_1 are frequencies lying about it at which the response is 3 db down. Hence $f_2 - f_1 = \Delta f$ is the 3-db bandwidth of the circuit.

Also,
$$Q = \frac{f_0}{\Delta f} = \frac{R_p}{\omega L} = \frac{\omega L}{R_s}, \quad (1)$$

where R_p is the parallel resistance, and R_s is the series resistance of the circuit.

From Eq. (1) it follows that Q is not only a function of bandwidth, but, for a given bandwidth, is proportional to the mid-band frequency.

The amplitude ratio, r, between two successive cycles of an exponentially damped wave is [1]:

$$r = \varepsilon^{-d}, \quad (2)$$

where d is the decrement of the circuit, and

$$d = \frac{R}{2 f_n L}, \quad (3)$$

where f_n is the natural frequency of the circuit.

One may write:
$$d = \frac{\pi}{Q}, \quad (4)$$

where Q is greater than 5.

Given two circuits of equal bandwidth, one resonant at f_1, the other at $f_2 = 2f_1$, then $Q_2 = 2Q_1$. Let a unit impulse be applied to each. Then an exponentially damped wave will arise from each circuit.

From Eq. (4), $d_1 = \frac{\pi}{Q_1}$ and $d_2 = \frac{\pi}{2Q_1}$.

Then $r_1 = \varepsilon^{-\frac{\pi}{Q_1}}$ and $r_2 = \varepsilon^{-\frac{\pi}{2Q_1}}$.

Now let $T = \frac{1}{f_2}$, and, as $f_2 = 2f_1$, $\frac{1}{f_1} = 2T$.

The decay in the amplitude of an exponentially damped wave over a period of 2T is r_1 at frequency f_1. At frequency f_2 it becomes r_2^2, as two periods occur during the time interval 2T.

However, $r_2^2 = \left[\varepsilon^{-\frac{\pi}{2Q_1}}\right]^2 = \varepsilon^{-\frac{2\pi}{2Q_1}} = \varepsilon^{-\frac{\pi}{Q_1}} = r_1$, and hence the decay over the time interval 2T is independent of frequency.

EFFECT OF PULSE FALL TIME

Evidently one must consider the relative response of the input circuit to the driving impulse to explain the observed results.

Figs. 2 (a) and (b) are oscillograms of the rectified radio-frequency pulse from the mixer crystal with a 330-ohm termination. The amplitude is about 2 volts. The trailing edge, which causes secondary shock effects, has a fall time of 1.5×10^{-2} microseconds, as measured in Fig. 2 (b). Table II shows the fall time expressed as a percentage of the period $T = 1/f$.

TABLE II

SHOCK IMPULSE AS A FUNCTION OF $1/f = T$

f (Mc/s)	T (sec)	Percent of T
30	3.33×10^{-8}	45.3
60	1.66×10^{-8}	90.5
120	8.35×10^{-9}	180.0

The much lower secondary shock observed at 60 Mc/s is thus explained, and one would expect an extremely low response at 120 Mc/s where the impulse extends over nearly two complete periods. The masking effect should be very small when $\tau_a \geq \frac{1}{f}$, where τ_a is the fall time of the pulse. The desirability of choosing a higher intermediate frequency for pulses having fast rise and fall times is also evident.

REFERENCE

1. "Radio Engineering", Sandeman, p. 167, 2nd Ed. Rev., Chapman and Hall.

ACKNOWLEDGMENT

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