THE DISPATCHER: A NEW MESOBAND HIGH POWER RADIATOR

Carl E. Baum

Air Force Research Laboratory, Directed Energy Directorate
3550 Aberdeen Avenue SE, Kirtland AFB, NM 87110-5776, USA
Magdalena.lopez@kirtland.af.mil

ABSTRACT

For a switched oscillator let us consider some shock-excited resonant structure. This excitation may come in the form of the discharge of some slowly charged structure. This excitation may come in the form of the discharge of some slowly charged transmission-line capacitance through a fast-closing switch. This paper introduces some concepts for antennas appropriate to be used with switched oscillators to effectively radiate the oscillating waveforms. The first kind is a dipole-fed reflector in which the quarter-wave (in the dielectric medium) oscillator is connected to an antenna of approximately a half-wave long (in air or SF_6). The second kind of antenna is a TEM-fed reflector with blocking capacitor. It has a TEM feed of characteristic impedance Z_f, typically using two feed arms giving an impedance in the 100 Ω ballpark for a half reflector with ground plane.

INTRODUCTION

There are various kinds of *hypoband* (narrowband) high-power sources (microwave tubes, or Zatrons) which may produce a hundred cycles or so in a pulse with center frequency in the range of some hundreds of MHz to a few GHz, a range of interest. There are also impulsive *hyperband* high-power radiators associated with impulse radiating antennas (IRAs) [8] with band ratios of the order of two decades. Let us consider a class of *mesoband* sources with frequency bandwidths somewhere between these two.

This paper summarizes the dispatcher concept which involves a switched oscillator as a high-power source which in turn drives an appropriate antenna. We consider two candidates, the electric-dipole-fed reflector and the TEM-fed reflector. Such combinations can be called oscillator reflector antennas (ORAs).

SWITCHED OSCILLATOR

Let the excitation come in the form of the discharge of some slowly charged capacitance through a fast-closing switch. As illustrated in Fig. 1 the oscillator is charged through some high impedance so as not to load the oscillator after the switch fires. Note that the switch is at the back end of a structure (away from the antenna) which we model as a length of transmission line of characteristic impedance Z_c and transit time t_t. With antenna impedance Z_a (assumed real here) and $Z_a >> Z_c$ for frequencies of interest this forms a quarter-wave resonator with

$$f = \frac{4}{t_t} \ , \quad \lambda = \frac{\text{v}}{f} \ , \quad \text{v} = \left[\mu_0 \varepsilon\right]^{-\frac{1}{2}} \equiv \text{propagation speed in oscillator dielectric medium} \qquad (1)$$

On closing the switch a wave (ideally a step function) of amplitude $-V_0$ propagates to the left (in Fig. 1), with nearly a $+1$ reflection coefficient. This gives a transient voltage doubling which is characteristic of a Blumlein and is advantageous in the present application. The reflection coefficient at the antenna is more accurately

$$\rho = \left[1 - \frac{Z_c}{Z_a}\right]\left[1 + \frac{Z_c}{Z_a}\right]^{1} \qquad (2)$$

Neglecting switch impedance (assumed zero) we have a geometric series with alternating signs, describing an exponential decay, the dominant frequency in this as in (1), but now more accurately described as a damped sinusoid. In N cycles the amplitude is reduced to $\rho^{2\text{N}}$. If we set this equal to e^{-1} we have

$$N = -\frac{1}{2\ell n(\rho)} \simeq \frac{1}{4}\frac{Z_a}{Z_c} \quad , \quad Q = \pi N \qquad (3)$$

As one can see there is a considerable advantage in making Z_c small, in line with the considerations in p[1]. There are various ways to construct the quarter-wave resonator with various dielectric media [2].

ELECTRIC-DIPOLE-FED REFLECTOR

Figure 2 shows a special kind of dipole-fed reflector. The quarter oscillator is connected to an antenna of approximately half-wave long (in air or SF_6), or other length as one may wish. In a simple (but approximate) transmission-line model the characteristic impedance Z_c of the oscillator section is much less than Z_a of the antenna section. Note that the switched oscillator is below a ground plane, a convenient place to locate pulse-power equipment without interfering with the antenna radiation.

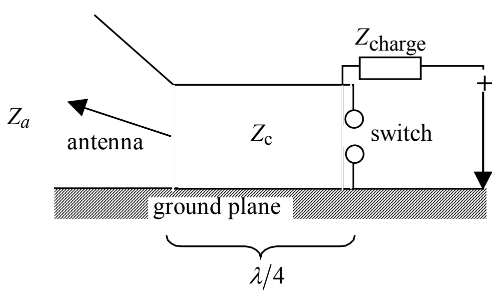

Fig. 1. Low-Impedance, Quarter-Wave Transmission-Line Oscillator Feeding High Impedance Antenna (Single-Ended Version).

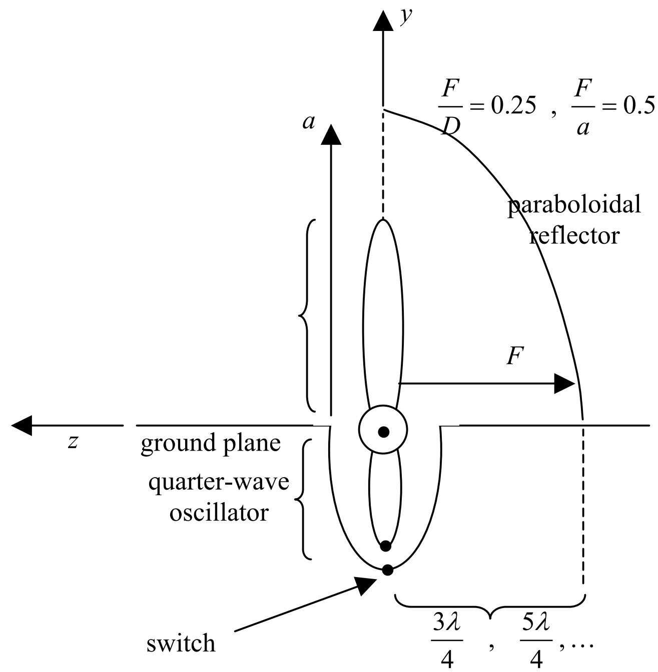

Fig. 2. Electric-Dipole-Ref Reflector.

For greater fields at a distance one needs some significant antenna gain. This is efficiently accomplished by a paraboloidal reflector with focus at the dipole-like element. Noting the polarization of electric-dipole fields the reflector should not extend to the left of the *xy* plane (or $z = 0$ plane), at least above the dipole (near the $+y$ axis). So one might choose

$$\frac{\text{focal length}}{\text{diameter}} \equiv \frac{F}{D} = 0.25 \ , \quad \frac{\text{focal length}}{\text{diameter}} \equiv \frac{F}{a} = 0.5 \tag{4}$$

Larger values intercept less of the power radiated by the dipole, so the above may be a reasonable choice. One could extend the reflector to $+z$ values near the ground plane, and truncate top portions of the reflector (near the $+y$ axis) which do not intercept much power. Consider now the size of the reflector as parameterized by the focal length F. Near the ground plane we can imagine an image dipole near $z = -2F$ with opposite current (due to negative reflection at the conductor reflector). In the forward direction (+z) of the reflector beam the forward radiation for the dipole will add to the reflector beam by judicious choice for F. To make the two waves add in phase in the near field we need

$$F = \frac{2n+1}{4}\lambda \ , \ \lambda = \frac{c}{f} = \text{wavelength} \ , \ \ c = [\mu_0 \varepsilon_0]^{-1/2} = \text{ speed of light} \ , \ n = \text{integer } \geq 0 \tag{5}$$

In the far field on the main beam the direct radiation from the dipole is small compared to that from the reflector.

Next, what is the optimal choice for n? Since $n = 0$ implies $a \simeq \lambda/2$ thereby touching the top of the dipole element (and giving only low gain). So let us constrain

$$n \geq 1 \ , \quad F = \frac{3\lambda}{4} \ , \ \frac{5\lambda}{4} \ , \ldots \tag{6}$$

A limitation of this type of antenna is the low radiation efficiency of a small antenna element. Note also that the antenna input impedance is not a simple constant resistive impedance. Further details are given in [3].

The there are the high voltage and switching considerations. In particular, the reflector can be part of a barrier to contain SF6. Near the base of the antenna element the insulting gas may not be adequate, and an oil region here may be appropriate.

TEM-FED REFLECTOR WITH BLOCKING CAPACITOR

Consider the antenna in Fig. 3. It has a TEM feed of characteristic impedance Z_f, typically using two feed arms giving an impedance in the 100 Ω ballpark. With the apex of this conical transmission line at the apex of a paraboloidal reflector, this has some similarity to the reflector IRA [4]. Since we are considering concentrating the operation around some radian frequency $\omega_0 = 2\pi f_0$ we can revisit the choice of the terminating impedance $\tilde{Z}_t(s)$ for optimum results. Note that $\tilde{Z}_t(s)$ is the parallel combination of the two terminating impedances connecting the two feed arms to the reflector. Here we use the two-sided Laplace transform with

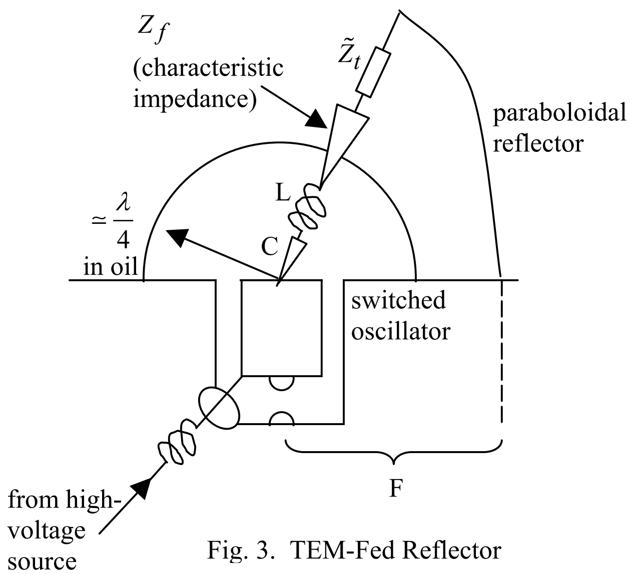

Fig. 3. TEM-Fed Reflector

$$\tilde{f}(s) = \int_{-\infty}^{\infty} f(t) e^{-st} dt \quad , \quad f(s) = \frac{1}{2\pi j} \int_{Br} \tilde{f}(t) e^{st} dt$$

$s \equiv \Omega + j\omega \equiv$ Laplace-transform variable or complex frequency (7)

$\gamma_d = \frac{s}{\text{v}}$, v = propagation speed in the dielectric medium of the switched oscillator

Near the output of the switched oscillator there is a blocking capacitance C (parallel combination of two in the arms). The charging current through C returns through the feed arms and reflector to the ground plane so $\tilde{Z}_t$ must be finite at low frequencies.

A disadvantage of a blocking capacitor is the impedance it presents to the oscillator, lowering the voltage delivered to the feed arms. One way to compensate for this is to include a series inductance L (net parallel combination) such that, at the frequency of interest, the series combination has zero impedance $\tilde{Z}_b(s)$, i.e.,

$$\tilde{Z}_b(s_b) = 0 = \frac{1}{s_b C} + s_b L = \frac{1}{j\omega_b C} + j\omega_b L_b \quad , \quad \omega_b = [LC]^{-\frac{1}{2}} \quad , \quad L = \left[\omega_b^2 C\right]^{-1} \tag{8}$$

Of course, a pulse contains a band of frequencies and the match is at what might be regarded as the center frequency. Note that series resistance can also be included in the above, if desired.

Modelling the switch closure by a voltage step $V_0 u(t)$ we need to optimize $V_a(t)$ in some sense as the waveform delivered to the antenna feed. (This can be added to the initial condition of $-V_0$ on the oscillator, if desired.) Defining a normalized impedance

$$\tilde{\zeta}(s) \equiv \left[Z_f + \tilde{Z}_b(s)\right] \Big/ Z_c \tag{9}$$

we have a reflection coefficient at the end of the oscillator (z = ℓ) as

$$\tilde{\xi}(s) \equiv \left[\tilde{\zeta}(s) - 1\right]\left[\tilde{\zeta}(s) - 1\right]^{1} \tag{10}$$

Due to shortness of space we skip over much of the mathematical details, which can be found in [3]. The voltage $\tilde{V}_a(s)$ on the antenna feed is just

$$\frac{\tilde{V}_a(s)e^{st_t}}{V_0} = \frac{2}{s}\frac{Z_f}{Z_c}\left[\left[\tilde{\zeta}(s)+1\right]+\left[\tilde{\zeta}(s)-1\right]e^{-2st_t}\right]^{-1} = \frac{2}{s}\left[\left[1+\frac{\tilde{Z}_b(s)+Z_c}{Z_f}\right]+\left[1+\frac{\tilde{Z}_b(s)-Z_c}{Z_f}\right]e^{-2st_t}\right]^{-1} \qquad (11)$$

Looking at the resonant behavior we need the zeros of the denominator. For the special case of $\tilde{Z}_b = 0$ we have the oscillator complex resonance as

$$-2s_0 t_t = \ell n\left(-\frac{1+\frac{Z_c}{Z_f}}{1-\frac{Z_c}{Z_f}}\right) \;,\; s_0 t_t = \left[\Omega_0 + j\omega_0\right]t_t = -\frac{1}{2}\ell n\left(-\frac{1+\frac{Z_c}{Z_f}}{1-\frac{Z_c}{Z_f}}\right) \pm j\frac{\pi}{2}$$
$$= -\frac{Z_c}{Z_f} \pm j\frac{\pi}{2} \text{ as } \frac{Z_c}{Z_f} \to 0 \;,\; f_0 = \frac{|\omega_0|}{2\pi} = \frac{1}{4t_t} \qquad (12)$$

If we modify $\tilde{Z}_b$ in (3.2) to give some series loss of an amount to make $\tilde{Z}_b(s_0) = 0$, this simplifies the analysis. Then the denominator in (3.12) is still exactly zero at the oscillator complex frequency. Defining

$$\chi \equiv \frac{2}{Z_f}\left[\frac{L}{C}\right]^{\frac{1}{2}} \qquad (13)$$

we find for the dominant pole pair in time domain

$$\frac{V_a(t)}{V_0} \simeq 2\left[\chi+\left[\pi\left[1-\frac{Z_c}{Z_f}\right]-\chi\right]\left[1+2\frac{Z_c}{Z_f}\right]\right]^{-1}\left[\frac{|\omega_0|e^{s_0[t-t_t]}}{s_0}+\frac{|\omega_0|e^{s_0^*[t-t_t]}}{s_0^*}\right]u(t-t_t)$$
$$\simeq 4\left[\pi+\frac{Z_c}{Z_f}[\pi-2\chi]\right]^{-1} e^{\Omega_0[t-t_0]}\sin\left(|\omega_0|[t-t_t]\right)u(t-t_t) \qquad (14)$$

For further insight, let $Z_c / Z_f \to 0$ giving

$$\Omega_0 \simeq 0 \;,\; \frac{V_a(t)}{V_0} \simeq \frac{4}{\pi}\sin\left(|\omega_0|[t-t_t]\right)u(t-t_t) \qquad (15)$$

which has amplitude $4/\pi$ which is the factor relating a square wave oscillating between +1 and –1 (or equivalently +2 and 0) to the fundamental sinusoidal component

For the case that there is an oil section at the beginning of the antenna feed this can be modeled as a section of transmission line, extending the analysis as in [5].

CONCLUDING REMARKS

Now we have design concepts for a new kind of high-power radiator. While the designs here are discussed in the context of a single-ended source in a half-IRA geometry, differential versions are also possible using a “full” reflector [6].

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